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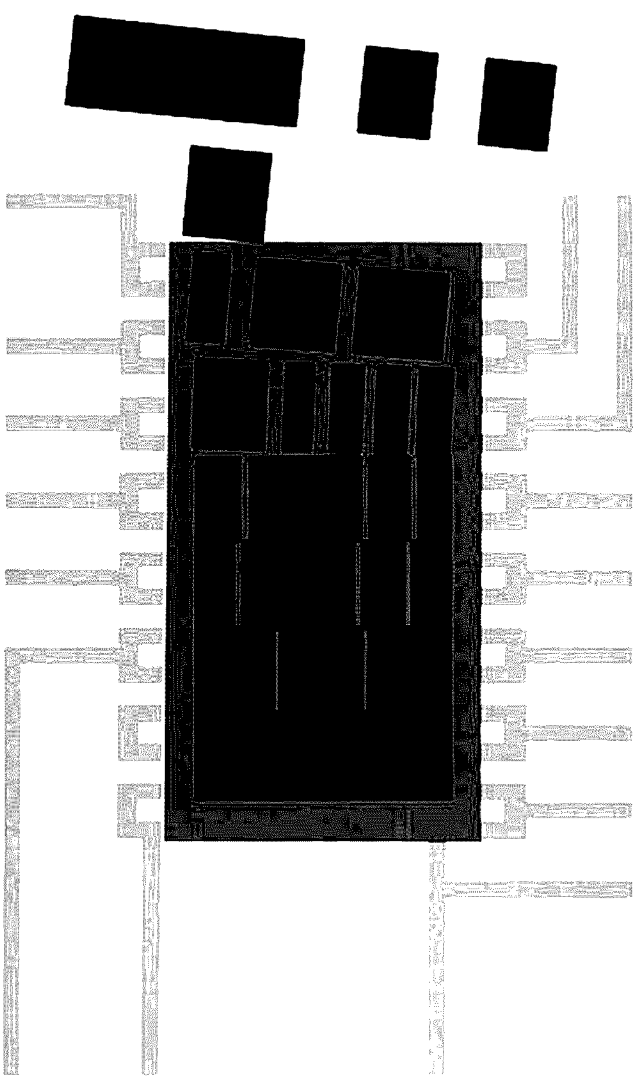
JANUARY 1974

cw memory for RTTY identification



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Hilda M. Wetherbee
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offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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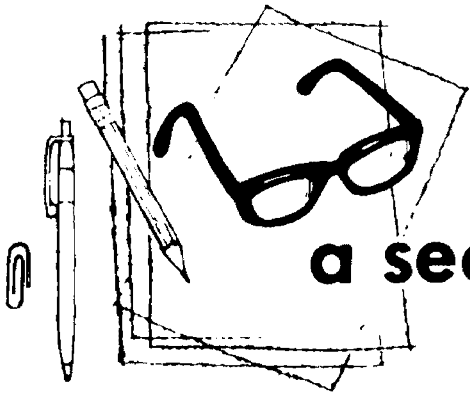


a second look

by jim fisk

More in '74. If you haven't seen that motto or heard of the concept, you will, because in 1974 *ham radio* will be providing a whole new family of services to the amateur community, services that will help each amateur derive the most enjoyment from his hobby.

More in '74 means a lot of things to *ham radio* readers, including more editorial staff and more specialized publications as well as some other exciting new projects. Since some of these projects are still in the embryonic planning stages we can't tell you much about them, because the final product may be considerably different than the original concept, but complete details will be announced as soon as they are available and those announcements will be well worth waiting for.

More in '74, among other things, means a new editor for *ham radio* and a new role for me as editor-in-chief. However, a change in editorial staff does not mean a shift in editorial policy. *Ham radio* will continue to do what we do best, bringing you the latest and best in technical articles and construction projects each and every month. The new editor, Joe Schroeder, W9JUV, has been an active amateur for more than 25 years and is well known to many of you. He's been associated with the electronics industry in one way or another for 20 years, was the editor of *Instrument Digest* and, more lately, editor of *Guns Illustrated.* Until leaving Chicago recently, he was on the Technical Committee of the Illinois Repeater Council. Joe is also an Honor Roll DXer with more than 340 countries to his credit and is active on all bands from 160 meters to 450 MHz. You'll be hearing a lot more from him in the future. Between the two of us *ham radio* will be bigger and better than ever.

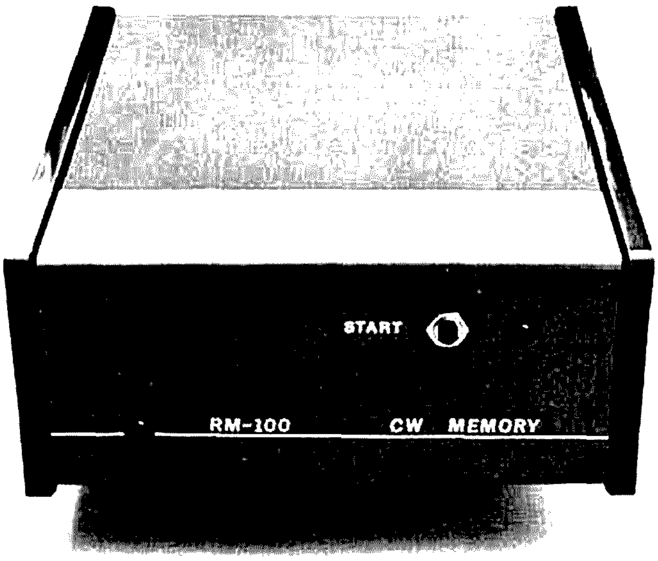
More in '74 means a brand-new newsletter, *HR Report,* which will keep you up to date with late-breaking news from the FCC, ARRL and industry sources, new DX activities, contest and hamfest announcements and up-to-the-minute propagation forecasts. The first issue of *HR Report* will be available in early January and will be sent out to subscribers twice monthly via airmail after that. In addition, special issues will be published as events warrant. If you want to know what's happening in the amateur world, you owe it to yourself to subscribe to *HR Report.* The subscription rate in the United States and Canada is $12.00 per year ($15.00 for overseas readers) with a guaranteed minimum of 24 issues per year.

More in '74 means a number of other new publications including new titles for your bookshelf and new operating aids for your station. For example, a new *Novice Radio Guide* is presently in production and will be available within a few months. Volume II of the popular *Ham Notebook* is currently in preparation and will be available later this year as will several other new titles.

These are only a few of the highlights — more announcements will be made as we progress through the year.

Ham Radio . . . more in '74.

Jim Fisk, W1DTY
editor-in-chief



CW memory

for RTTY identification

Complete construction details for the RM-100 — a 256-bit CW memory using modern read-only memory ICs

Howard L. Nurse, W6LLO, 665 Maybell Avenue, Palo Alto, California 94306

Do you find it frustrating to stop typing in the middle of an RTTY QSO to comply with FCC rule 97.87? This rule requires that you, as an RTTY station operator, keep your CW key at the ready to identify properly. The RATTline RM-100 CW memory is designed to facilitate that required CW identification.

The RM-100 uses an integrated-circuit memory to allow automatic CW transmission of a message at the speed of approximately ten words per minute. It can be connected directly to the RY-170 AFSK generator described in the December, 1973, issue of *ham radio* (see **fig. 1**), or to any other keying line compatible with the npn open-collector output of the RM-100.

When you want to identify, a CW message such as "... de W6LLO" can be started locally with a front-panel switch, or remotely with a ground control signal. While the RM-100 is keying, an LED lights to indicate the circuit is busy.

This article shows how to construct the RM-100 CW Memory. Instructions are given for programming your own memory, or you can purchase one already programmed.* The RM-100 circuit can be

*JTM Associates, P.O. Box 843, Manchester, Missouri 63011 ($12.50); Babylon Electronics, Carmichael, California 95068 ($15.00).

built for under twenty-five dollars using the pre-programmed memory, or for under twenty if you program the memory yourself.

programming the memory

At the heart of any automatic message generator is a means to store binary information. In the past, code wheels with cogs, or diode matrices, have been popular as memories. A relatively new type of memory is now available to amateurs from surplus suppliers. The programmable-read-only memory integrated circuit, abbreviated as PROM, P.ROM, or fPROM, depending on the respective manufacturer, can be custom programmed in the field by applying the correct voltages to its inputs, following the manufacturer's instructions.

The Signetics 8223 Programmable ROM IC I chose for the RM-100 is the easiest to find.* The 8223 is supplied with all of its 256 memory cells at a TTL logic low or 0. The memory is programmed by fusing (burning out) microscopic wires in selected cells to generate a desired pattern of logic highs, or 1s. Once the memory has been programmed, the stored binary information can be recalled from each cell by applying an address word to the inputs of the device.

*JTM Associates, *previously foot-noted, or* Poly Paks, P.O. Box 942, Lynnfield, Massachusetts 01940 ($7.95).

The first step when programming is to decide what you want the memory to say. A Morse code *key-down* is defined as 1, while a *key-up* is 0. Using a pattern as shown in fig. 2, mark each of the eight rows of 32 memory cells where a Morse code *key-down* is desired. Remember that a dot will occupy one memory cell, a dash or letter space, three, and a word space, four. Ten consecutive *key-up* cells indicate an "end-of-message." **Table 1** lists the letters of the alphabet and numbers from 0 to 9 with the required number of memory cells for each character. The CW message can occupy 246 memory cells after the 10-bit allowance has been made for the end-of-message code.

Once you have established the program matrix, the following procedure should be used with the circuit shown in

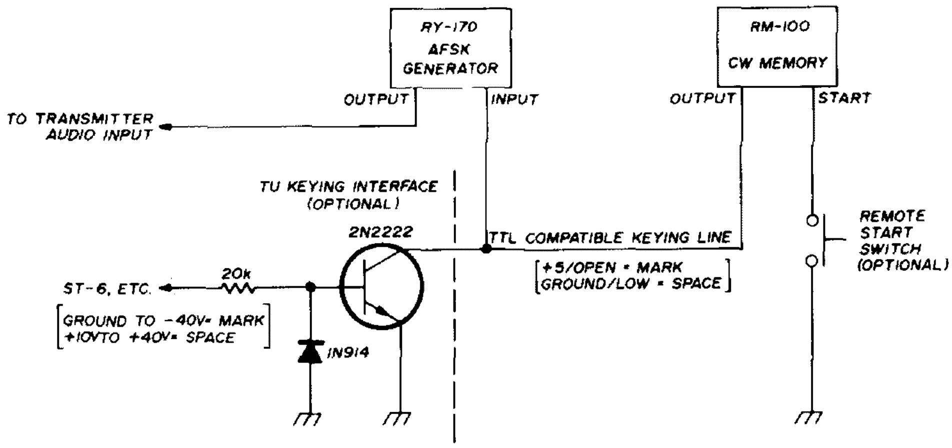


fig. 1. RATTline system interconnection showing RY-170 AFSK generator and RM-100 CW memory wired together with one possible interface circuit to a ST-6 RTTY terminal unit.

table 1. Memory cells required for letters, numbers and punctuation.

A 8	H 10	O 14	V 12
B 12	I 6	P 14	W 12
C 14	J 16	Q 16	X 14
D 10	K 12	R 10	Y 16
E 4	L 12	S 8	Z 14
F 12	M 10	T 6	
G 12	N 8	U 10	
1 20	6 14		, 22
2 18	7 16		? 18
3 16	8 18		. 20
4 14	9 20		/ 16
5 12	Ø 22	end of msg	10
		word space	4

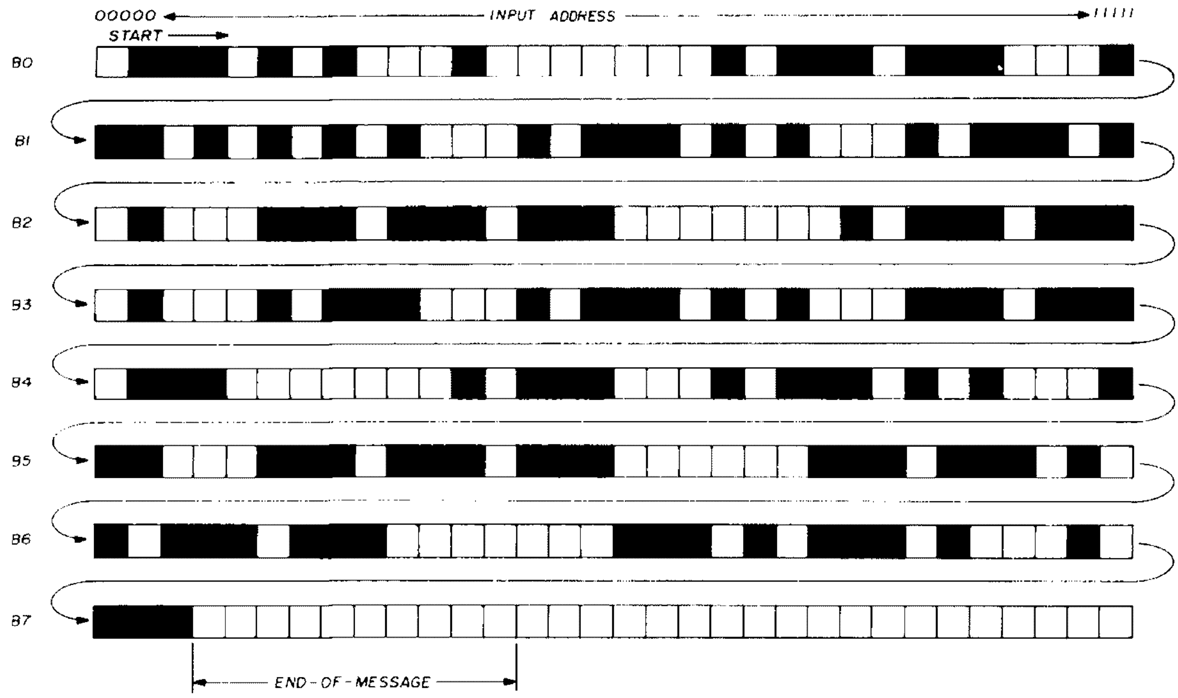


fig. 2. CW programming matrix consisting of eight rows of thirty-two memory cells, for a total of 256. Message shown here is "de W6LLO Palo Alto, Ca."

fig. 3 to program your memory.* Be careful – you only have one chance to correctly program each cell. Once the wire in a cell has been fused, it cannot be returned to its 0 logic state. If changes are required in the message, it is necessary to program a replacement integrated circuit.

programming procedure

The 8223 ROM IC is shipped with all outputs at logical 0. To write a logical 1 proceed as follows:

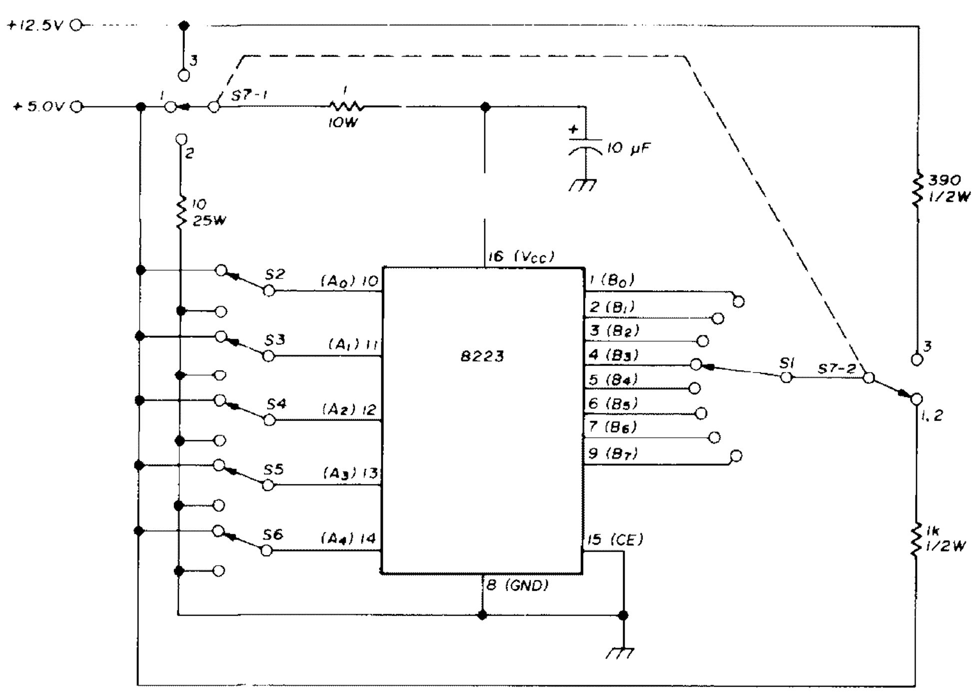
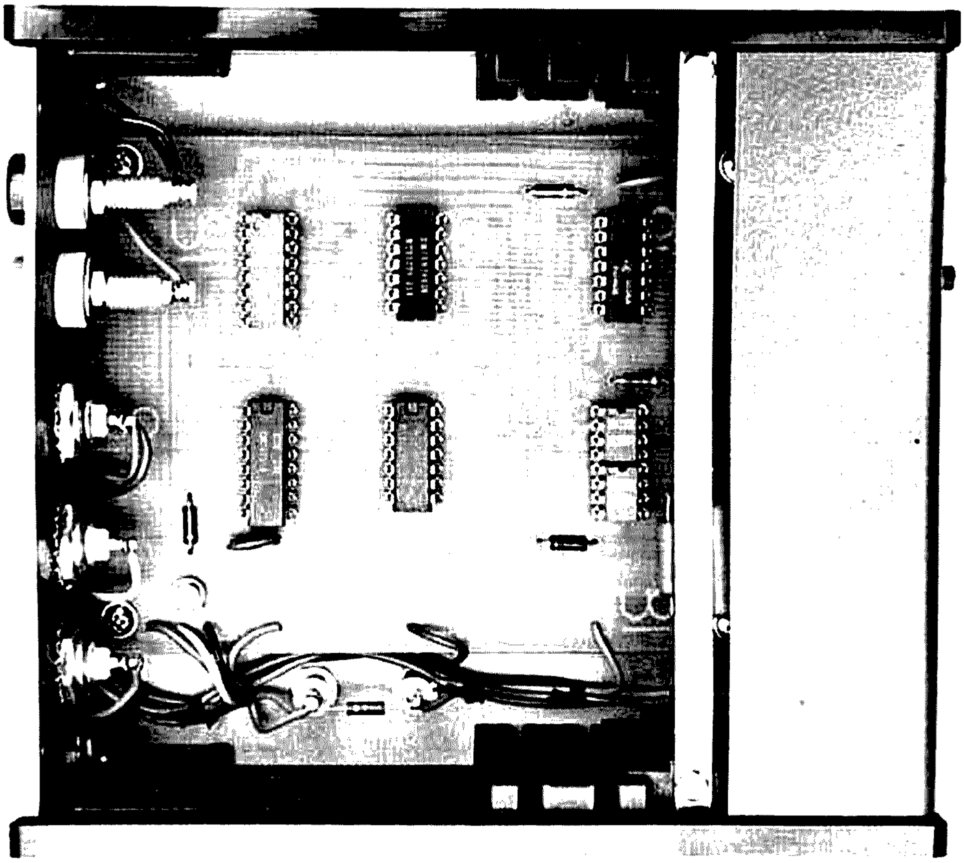


fig. 3. Programmer schematic for Signetics 8223 field-programmable read-only memory. The 10-μF capacitor from pin 16 to ground is required to eliminate noise from the supply line. During programming switch S7 must be in position 2 long enough for the 1-μF capacitor to discharge to less than 0.5 volt.

S1	single-pole, 9-position switch
S7	2P3T rotary switch with ground connected to the middle position of the first section. As V_{CC} (pin 16) is taken from 5 volts to 12.5 volts it will momentarily go to ground.



Inside of RM-100 as viewed from top. Resistor mounted on stand-offs is for LED indicator light.

1. Start with pin 8 grounded and V_{cc} removed from pin 16.

2. Remove any load from the outputs.

3. Ground the Chip Enable.

4. Address the desired location by applying ground (i.e., 0.4 V maximum) for a 0, and +5.0 V (i.e., +2.8 V minimum) for a 1 at the address input lines.

5. Apply +12.5 V ±0.5 V to the output to be programmed through a 390-ohm, 10% resistor. Program one output at a time.

6. Apply +12.5 V to V_{cc} (pin 16 for 50 milliseconds to 1 second (maximum) with a V_{cc} risetime of 50 microseconds or less. If 1.0 second is exceeded, the duty cycle should be limited to a maximum of 25%. The V_{cc} overshoot should be limited to 1.0 V maximum. If necessary, a clamping circuit should be used. The V_{cc} current requirement is 40 mA maximum at +12.5 V. Several fuses can be programmed in sequence until 1.0 second of high V_{cc} time is accumulated before imposing the duty cycle restriction.

Note: Normal practice in text fixture

*Programming instructions and fig. 3 adapted from Signetics Catalog, 1972, page 4-10.

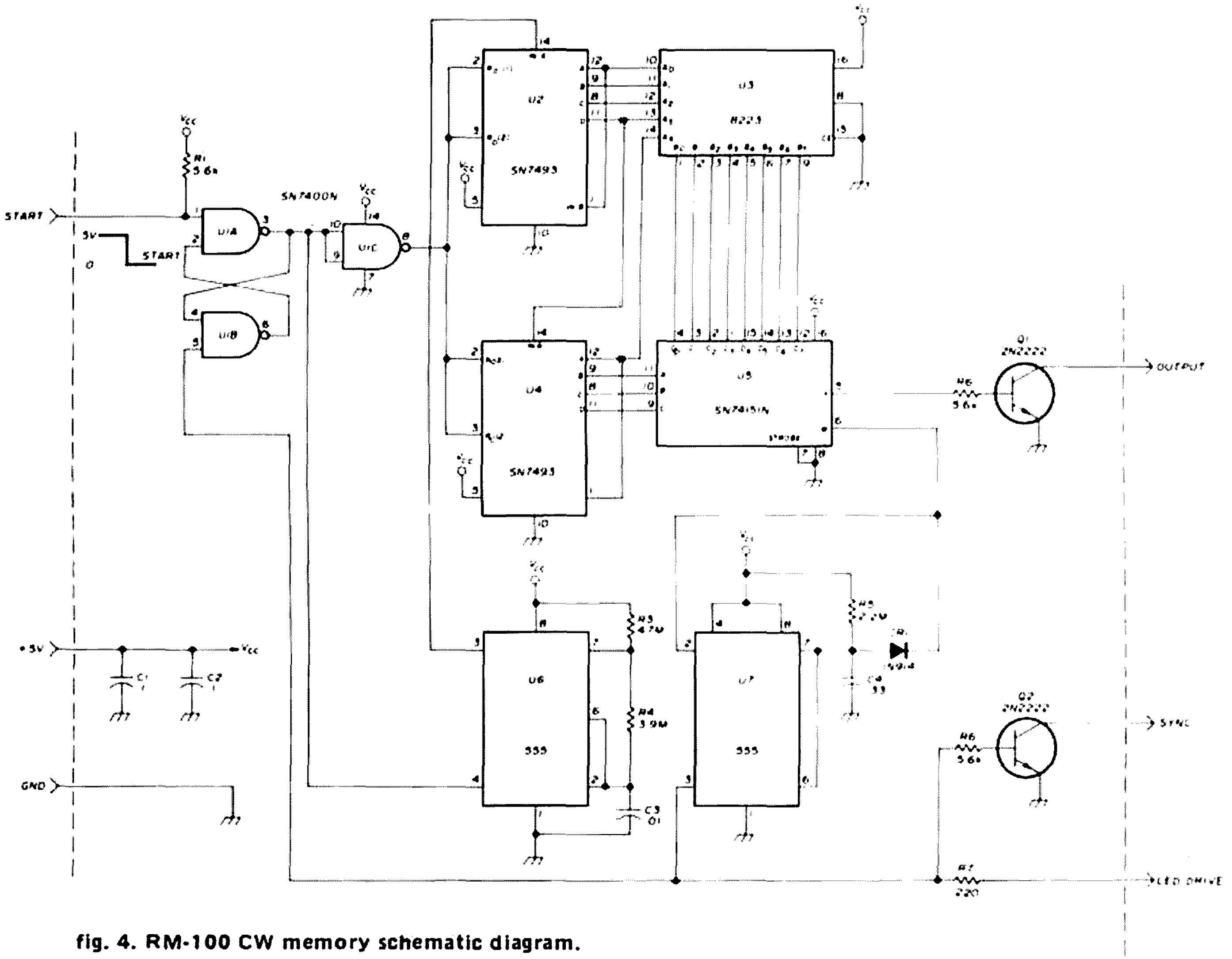


fig. 4. RM-100 CW memory schematic diagram.

layout should be followed. Lead lengths, particularly to the power supply, should be as short as possible. A capacitor of 10 μF minimum, connected from +12.5 V to ground, should be located close to the unit being programmed.

7. Remove the programming voltage from pin 16.

8. Open the output.

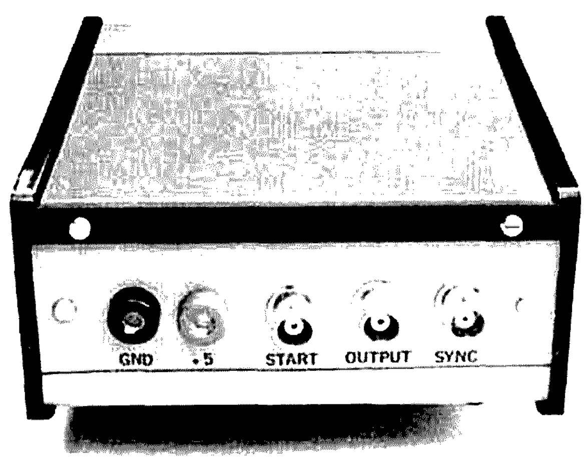
9. Proceed to the next output and repeat, or change address and repeat procedure.

10. Continue until the entire bit pattern is programmed into your custom 8223 ROM IC.

By now you have obtained a programmed memory, either by programming it yourself, or by purchasing it pre-programmed. Now you are ready to use it in the RM-100 circuit.

circuit description

The RM-100 CW memory consists of clock, divider, memory, data selector and end-of-message sense circuitry. The unit schematic is shown in fig. **4**, and the component parts layout is given in fig. **5**.



Rear panel of the RM-100.

Photographs of the completed RM-100 appear throughout this article.

The clock frequency established by timer U6, R3, R4 and C3 determines the dot length of the CW output. The code speed which results from a Morse dot length of 88 milliseconds was chosen, on suggestion from W6FFC, to keep the receiving RTTY machine in synchronism

Timer U7 detects when ten cells have passed without a keying event. Keying dots or dashes from the data selector discharges timing capacitor C4, whereas the timer is triggered on the trailing edge of each dot or dash. If a new keying bit does not discharge the capacitor within 880 milliseconds after the timer has been triggered, the timer output will reset the

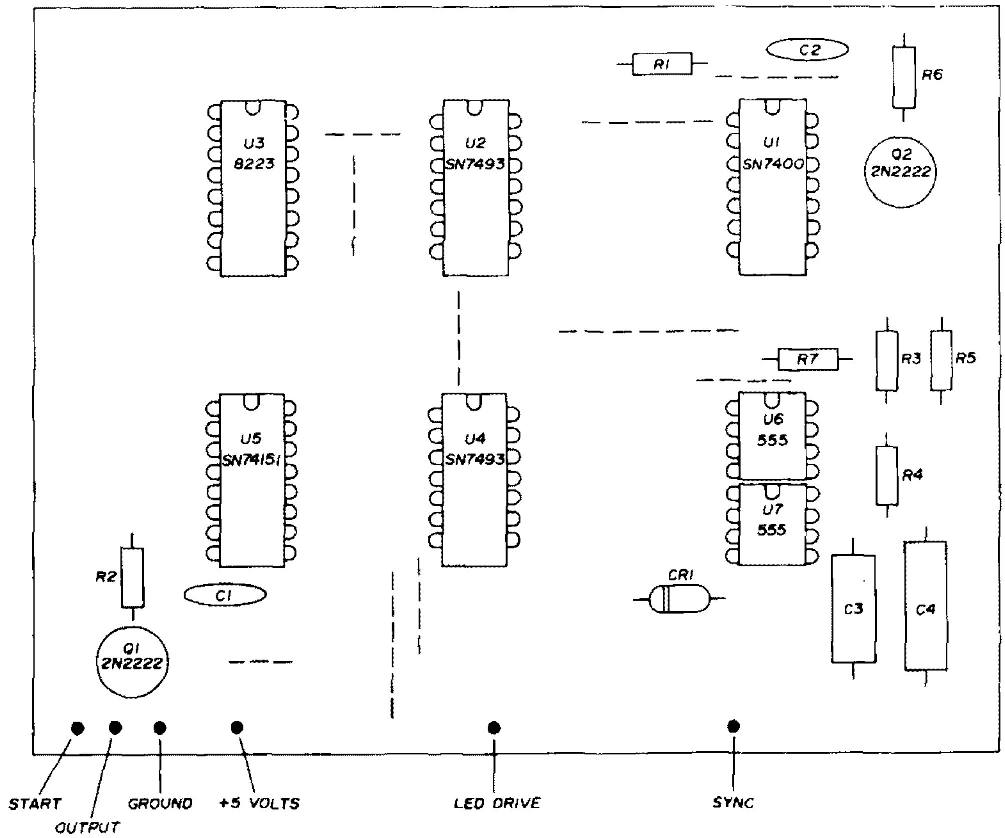


fig. 5. Suggested component layout for the RM-100. Dotted lines denote jumpers.

with the incoming pulses. Instead of printing garble, the machine will print blanks and Os in a combination which depends on the particular letter being sent.

The clock output frequency is divided by binary counters U2 and U4. The first five outputs from the dividers supply counts 0 through 31 to address the memory, while the last three outputs are used to address the data selector, U5.

The memory, U3, is read out through the data selector, one cell at a time, in the order shown in **fig. 2**. The Q output from the selector drives the base of keying transistor, Q1, while the $\overline{Q}$ output is used by the end-of-message circuit.

control flip-flop U1A and U1B, and the counters. The RM-100 CW message can be started again after the counters have been reset.

The sync output from the memory is connected to an open-collector transistor switch which turns on for the duration of a message. The sync output will be used by upcoming RATTline accessories.

As with the RY-170, a separate supply is required to power the RM-100. A 5 ±0.25 Vdc regulated supply capable of 150 mA should be used.

construction

The RM-100 can be built using perforated or printed-circuit board. The cir-

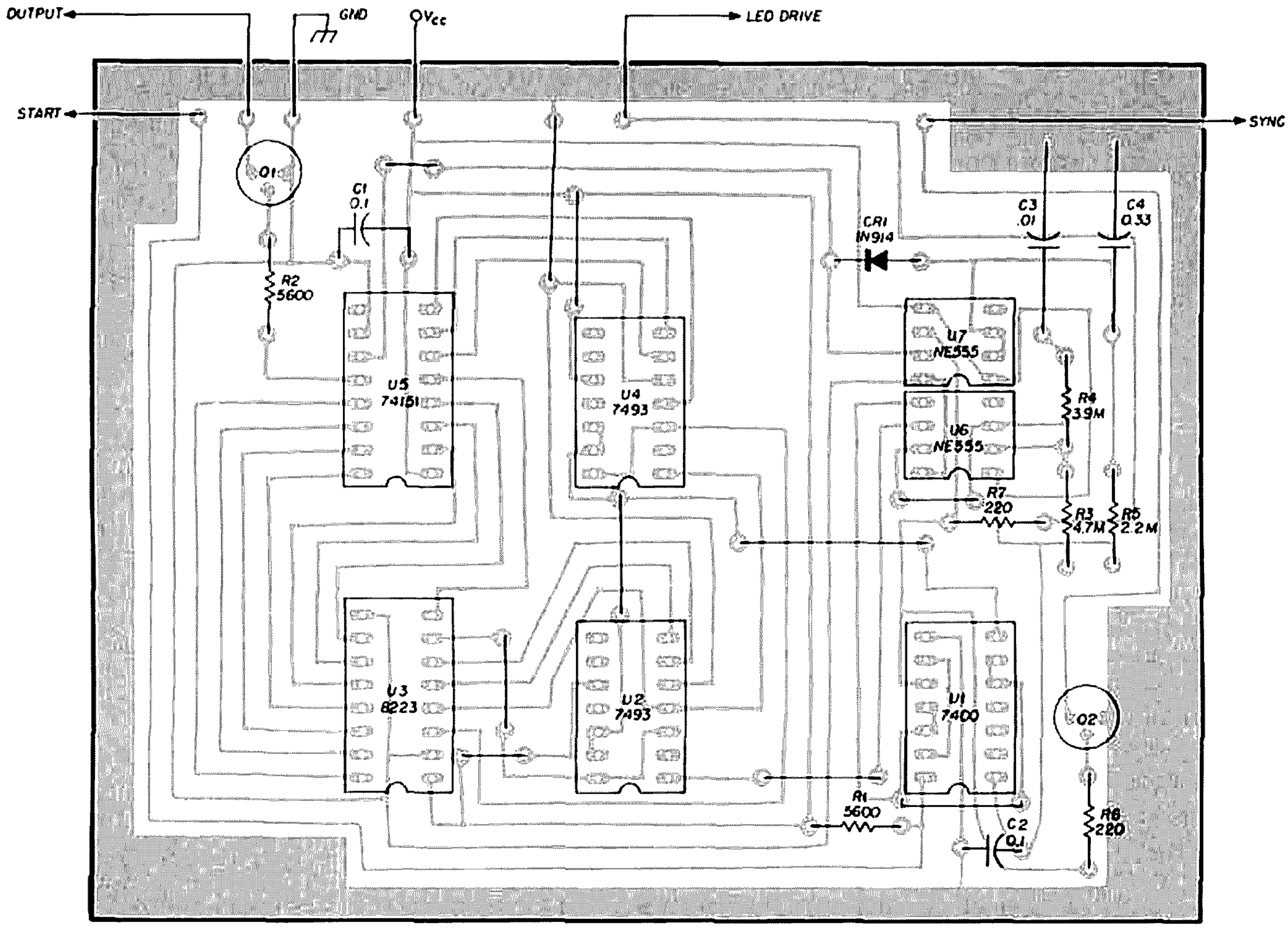
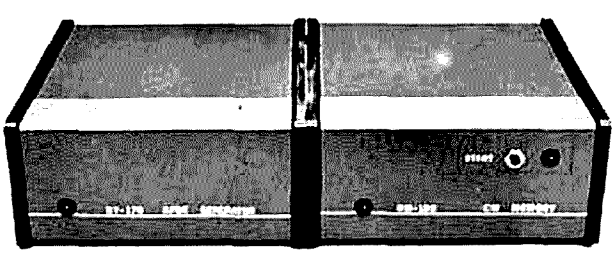


fig. 6. Full-size printed-circuit layout (foil side) for the 256-bit CW memory.

cuit should be placed in an enclosure, such as the Ten-Tec JG-5 shown in the photographs, to shield against rf.

An LED, powered from the 5-volt supply through a 220-ohm resistor, serves as a pilot light. A second LED is turned on by U7 when a message is being transmitted. A pushbutton switch on the front panel, which grounds the *start* control line, is used to trigger the memory.



RM-100 CW memory alongside the RY-170 AFSK generator.

The only critical components are the capacitors and diode used in the timers. High stability, low leakage capacitors should be used at C3 and C4 to maintain correct timing. A silicon diode is required at CR1 so that diode leakage will not effect timing in the end-of-message circuit.

Bypass capacitors should be used in at least two places on the circuit board. They should be near integrated circuits from V_{cc} to ground to filter transients. Molex pins hold the integrated circuits, and BNC jacks are used to connect signal inputs and outputs. Shielded cables are recommended when connecting the RM-100 with other parts of your system to reduce the possibility that strong rf fields will interfere with circuit operation.

Don't let rule 97.87 get to you! Now you can have automatic CW identification in your RTTY system with the RM-100 CW Memory.

ham radio

five-band kilowatt linear

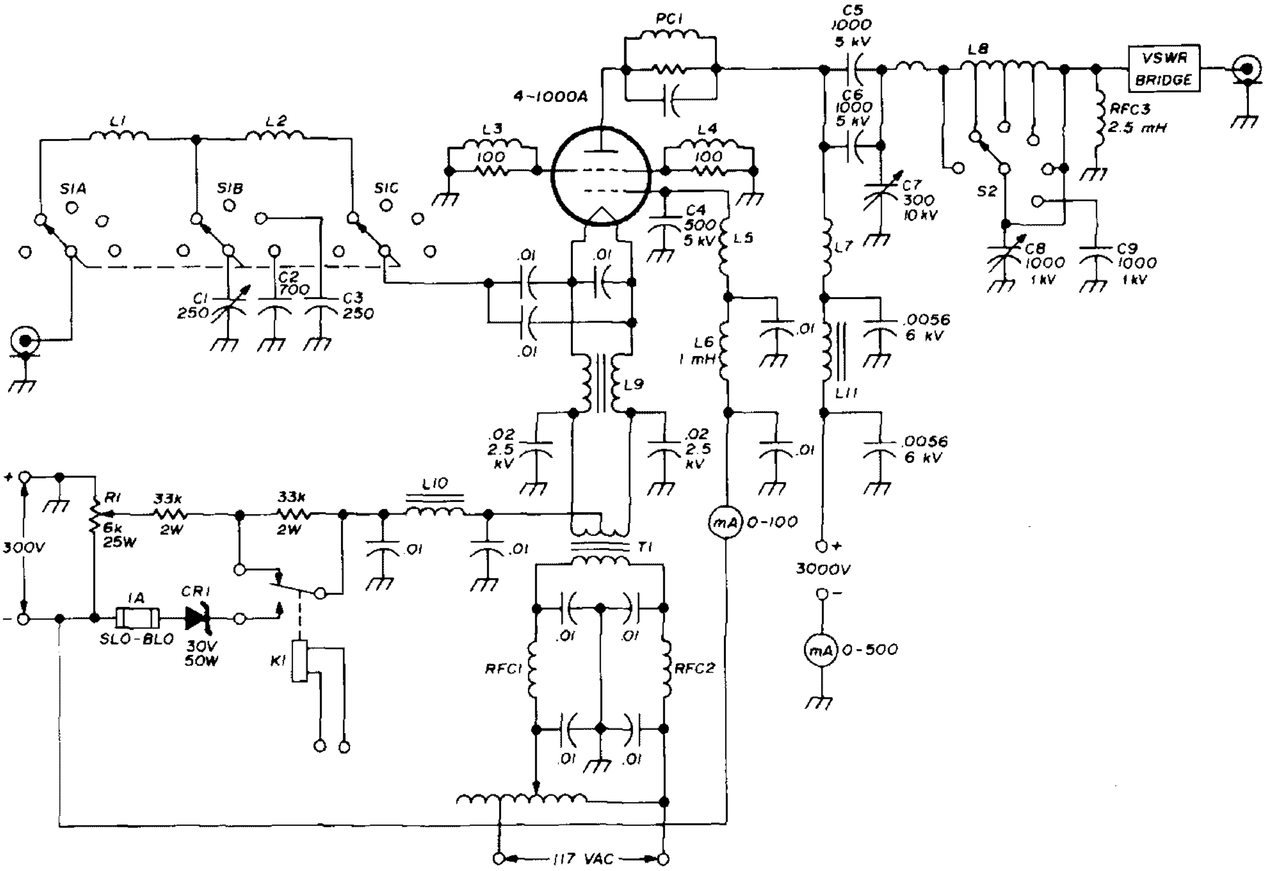
How to use beam power tetrodes efficiently and economically in the grounded-grid, grounded-screen configuration

John R. True, W4OQ, 10322 Georgetown Pike, Great Falls, Virginia 22066

With the proper design precautions, a power tetrode in a grounded-grid and grounded-screen configuration can provide more than 13-dB stable, parasitic free gain in the high-frequency spectrum. An old 4-1000A which I use in this circuit (see **fig. 1**) operates at 2-kW PEP input when driven by an exciter rated at 100-watts output. Since power tetrodes have very high power gain characteristics, they require very little drive. Typically, the amplification factor of the grid is in the order of five to six times that of the screen. However, grid dissipation is very low and screen dissipation is considerably greater.

For amateur use, however, the high gain characteristic of the power tetrode is not always an advantage, because few amateurs use exciters that operate efficiently at less than 100-watts output. When the 100-watt exciter is used to drive a high-gain power amplifier, it must be loaded with a power attenuator, which wastes valuable power. One commercial linear recently offered on the market, for example, required a power-wasting 12-dB pad.

The high gain of the power tetrode can also lead to other problems, such as vhf parasitics which require extensive suppression. Also, a tetrode, grounded-filament, class-AB2 linear amplifier re-



C1	25-250 pF air variable, 1000-volt working
C4	500 pF ceramic, 5 kV working (Centralab 858S)
C5,C6,C9	1000 pF ceramic, 5 kV working (Centralab 858S)
C7	10-300 pF vacuum variable, 10 kV
C8	1000 pF air variable, 1000 V
L1,L2	see table 1
L3,L4	3 turns no. 12, 3/8" diameter, with 100-ohm, 2-watt resistor
L5,L6	1 mH rf choke
L7	plate rf choke, 1 amp (B&W model 800)
L8	plate tank assembly (B&W 850A)
L9	33 turns no. 10 enamelled bifilar wound on ½" ferrite rod, 7" long
L10,L11	2½" no. 18 wire wound around ½" ferrite rod
K1	spdt relay, 117-Vac coil
PC1	20 pF ceramic capacitor (Centralab 852S) in parallel with four 470-ohm, 2-watt resistors and 4 turns no. 12 ½" diameter, resonated to 110 MHz
RFC1,RFC2	30 turns no. 16, closewound on ½" diameter wooden dowel
S1	3-pole, 7-position ceramic switch
S2	1-pole, 6 position switch (part of B&W 850A)
T1	filament transformer, 7.5 volts, 21 amps

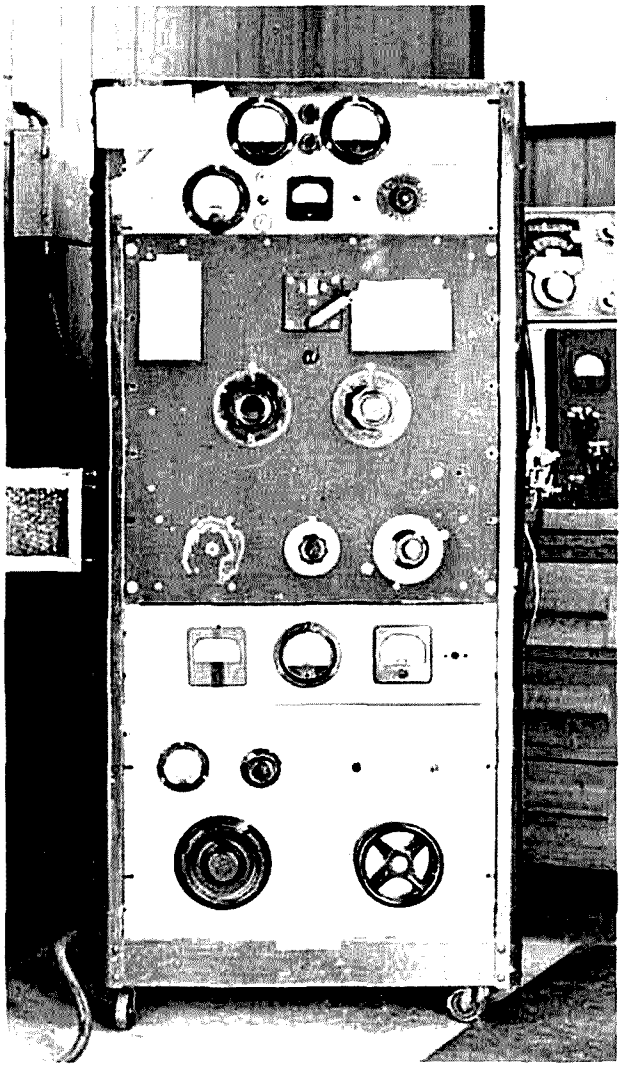
fig. 1. Modified grounded-grid, grounded-screen, beam-tetrode power amplifier. Plate and screen power supplies are shown in fig. 3. Grid bias is provided by the zener diode, CR1.

quires a well regulated grid and screen supply. Using two tubes in parallel just multiplies the problems.

In recent years the grounded-grid triode has been used extensively as a linear amplifier for ssb and CW. Among its many advantages are excellent input/output circuit isolation (which means less parasitic problems), high power gains (10 to 13 dB), possibility of fixed-tuned bandpass input circuitry and no requirement for a regulated screen supply (or any screen supply at all, for that matter). Using the grounded-grid circuit, it is very easy to build a stable linear amplifier with low distortion products which is relatively free from harmonic distortion.

With all the advantages of grounded-grid triode power amplifiers, why even bother with a tetrode? Primarily because

tetrodes are in abundant supply at ridiculously low prices — the broadcast industry practically gives them away after a certain number of operating hours. This is less expensive for them than experiencing a failure in the middle of an important broadcast. There are also a number of



Rack-mounted linear amplifier includes meters for vswr, grid and plate current, and filament, grid, screen and plate voltage. Variac-controlled power supply is located behind bottom panel.

power tetrodes available on the surplus market. However, if you are selecting a tetrode for a linear amplifier, choose *one* with adequate plate dissipation and filament emission to get the job done properly.

Experience has taught me that all is not gold that glitters. If the tetrode is used with the grid and screen grounded, the power gain will be low due to degenerative feedthrough power. Additionally, the grid hogs control of the space current while the screen loafs. To drive the plate to full output, excessively high grid current is required.

In an effort to cure some of these shortcomings, I built an amplifier with the grid less than fully grounded. It was bypassed with a 500-pF capacitor and the dc brought out for metering through an rf choke. This simple technique reduced the grid current (and increased screen current) to an appropriate value, but power gain was reduced because the screen is not as effective as the grid in controlling plate current. There had to be a better way!

the circuit

Considering the tremendous gain possible with a conventional grounded-filament, class-AB2 tetrode amplifier, I thought a hybrid compromise might be the answer. That is, to retain the isolation provided by the partially grounded grid and grounded screen, but to place some voltage on the screen to increase transconductance. That is easily managed with a grounded screen by biasing the filament negative in respect to ground. Thus, the screen becomes positive with respect to the filament. Unfortunately, so does the grid! Placing a zener diode in series with the filament return to the minus side of the screen supply will provide the grid with well-regulated bias for satisfactory idling as well as full input plate current.

However, this is not the complete answer. If you look carefully at the physical structure of the grid and screen of the 4-1000A, you will see that they form a sort of basket-woven cavity. When a positive voltage is placed on the screen, these cavities (with the filament) act as a triode oscillator above 100 MHz. By way of demonstration, if you ground the grid and screen of a 4-1000A and couple a grid-dipper to these leads, you will find that the screen is resonant at about 120 MHz. The grid is resonant at about 110 MHz, an ideal condition for *regeneration.*

Under these conditions, the screen is inductive in respect to the grid, the condition required for in-phase feedback.

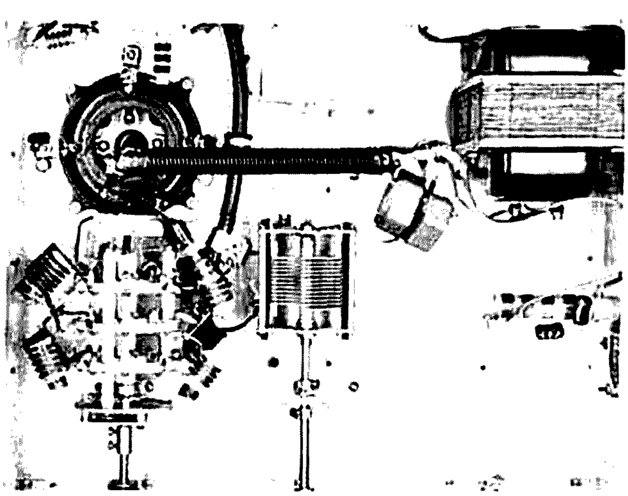
In fact, a 4-1000A self-oscillates near 110 MHz, and the 4-400A at about 140MHz. A pair of parallel power tetrodes may act like a push-pull oscillator, oscillating at a frequency that is dependent upon the length of their interconnecting leads.

If the grid is made to resonate higher (or the screen to resonate lower), self-neutralization occurs since the screen is capacitive in respect to the grid, producing *degeneration*. This is easily accomplished. As noted before, the small 500-pF bypass capacitor slightly raises the resonant frequency of the grid. A small inductor in series with the screen shifts its natural resonance point below that of the grid, resulting in neutralization.

power supply considerations

If you trace the plate-current path in fig. 1, you will find that the plate current flows through both the plate and screen supplies. Therefore, the screen supply must furnish both screen and plate current. The screen voltage must also be added to the plate voltage when calculating plate input power (subtract the zener voltage drop). It should be obvious, therefore, that the screen supply must provide very good voltage regulation.

Using the 4-1000A as an example, assuming a plate supply of 3000 volts, a screen supply of 300 volts and zener bias of 30 volts, the idling plate current will be about 125 mA. With 100-watts of excitation and a desired plate power input of 1 kilowatt, grid current is approximately 30 mA, screen current is about 100 mA, and plate current is approximately 300 mA.



Underneath the linear amplifier chassis. Input matching network is at lower left.

To reduce noise generation while receiving, some means must be provided for completely turning off the power tube

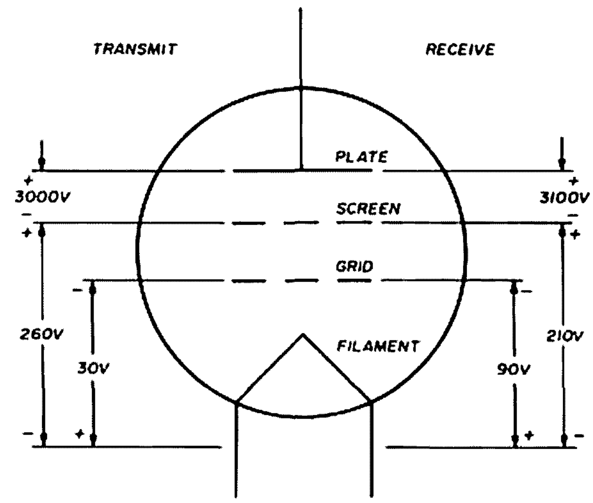


fig. 2. Operating voltages for the 4-1000A power amplifier during transmit and receive. Cutoff voltage on receive is provided through the contacts of relay K1 (see text).

when not transmitting. Fig. 1 shows the circuit I use to accomplish this. When transmitting (relay K1 picked up), the grid is minus with respect to the filament by the voltage drop of the zener diode. When the relay drops out (receive position), the filament assumes the positive voltage offset from a tap on the screen supply bleeder resistor. A grid-filament bias of −90 volts is more than enough to completely cut off a 4-1000A with the plate and screen voltages indicated in fig. 2.

A complete schematic of the plate and screen power supplies is shown in fig. 3. All supply voltages are brought up to full operating levels with a variable transformer (Powerstat). The large filter capacitors require a lot of current to load them to full voltage, and this procedure reduces the surge current through the silicon diodes. Another Powerstat is used with the filament transformer to reduce the thermal shock of an instant-on filament switch.

The plate and screen power supplies

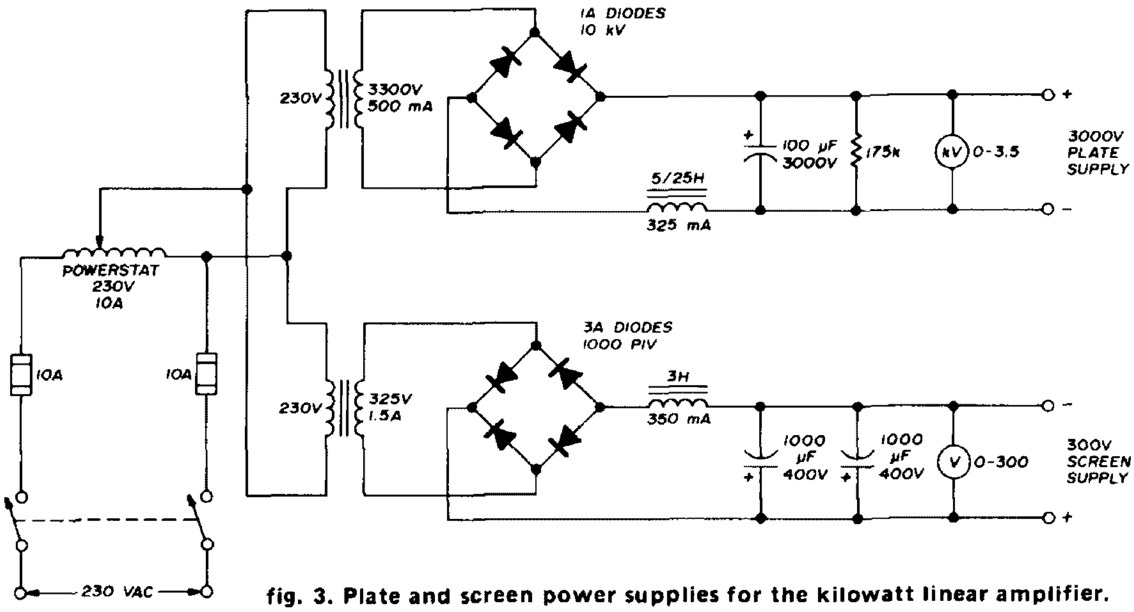


fig. 3. Plate and screen power supplies for the kilowatt linear amplifier.

use common protective fuses. This avoids the possibility of losing the plate supply with the screen supply still on, which would surely destroy the screen of the 4-1000A. The 1-amp slo-blo fuse between the grid-bias zener diode and the minus side of the screen supply protects against overloads, including parasitics. Any excessive dc cathode current will blow the fuse, automatically returning the grid-filament bias to the cutoff value.

input network

Since normal voice waveforms have an approximately 3.5:1 peak-to-average ratio, this must be considered when computing the input impedance to the grounded-grid stage. At 1-kW input total required filament emission is 430 mA (300 mA plate current + 100 mA screen current + 30 mA grid current). With a 3.5:1 peak-to-average voice ratio, peak filament emission is 1.5 amperes (3.5 x 430 mA). Therefore, filament input impedance is

$$Z_i = \frac{P_i}{I^2} = \frac{100}{(1.5)^2} = \frac{100}{2.25} \cong 45 \text{ ohms}$$

where Z_i is the input impedance, P_i is the input (drive) power (100 watts), and I is the peak filament current.

The graphical design of a T-network for matching 50-ohm coaxial cable to the 45-ohm input impedance of the 4-1000A is shown in fig. 4.[1] In this graphical solution, the transfer impedance (Z_T) was chosen to be 150 ohms. As can be seen, X_C = 51 ohms, X_{L1} = 71 ohms and X_{L2} = 69 ohms. The Q of this network is approximately 3. Practical network component values for each of the high-frequency amateur bands are listed in table 1.

The T-network inductors L1 and L2 are wound on ½-inch polyethylene tub-

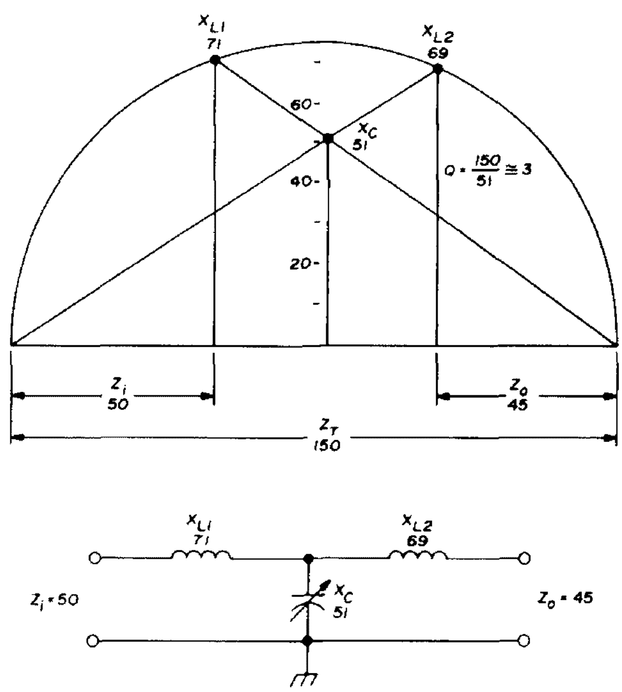


fig. 4. Graphical design of the T-network used at the input of the power amplifier.

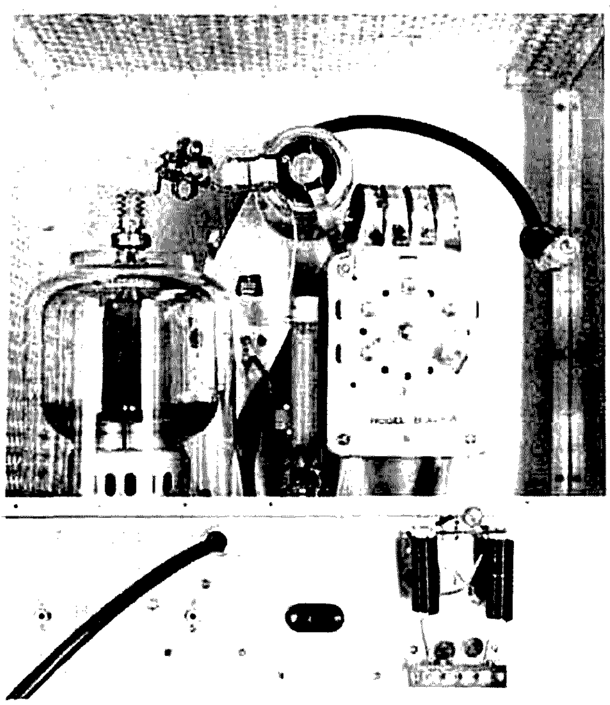
ing. Number-14 Formvar is used for the 80- and 40-meter coils while number-12 Formvar is used for 10, 15 and 20 meters. A ½-inch wooden dowel is inserted in to the tubing during winding to keep the tubing from being deformed.

After the coils are wound, a short length of ½-inch OD ferrite rod (about 1-inch long) is inserted into the tubing and adjusted for the proper inductance value given in table 1. By using 5% fixed mica capacitors across the inductors and a grid dipper to check for resonance, it is possible to adjust them very close to their required inductance values. When the inductors are completed, coil dope is used to hold the turns and the ferrite slug in place.

output network

The output pi network is based on the B&W model 850A bandswitching pi-network inductor which includes a built-in bandswitch. As shown in fig. 1, an additional switch contact was added to permit the use of a 1000-pF fixed capacitor in parallel with C8 on 75 meters.

The rear plate of the antenna loading capacitor (C8) is used as the common ground point for the output circuit. (The rear plate of the input capacitor, C1, is used as the common ground point for the input circuit). Both the input and output coaxial cable connections are made at their respective common ground point and switch. These cables are connected directly to the exciter and the vswr bridge without any terminations at the chassis. This reduces input-output coupling.



Rear view of output components of the linear showing the 4-1000A and B&W pi-network inductor.

table 1. Component values for input T-network described in fig. 3 (X_C = 51 ohms, X_{L1} = 71 ohms, X_{L2} = 69 ohms).

frequency	L1	L2	C
3.60 MHz	3.14 μH	3.05 μH	865 pF
3.90 MHz	2.90 μH	2.81 μH	800 pF
7.15 MHz	1.58 μH	1.54 μH	436 pF
14.20 MHz	0.80 μH	0.77 μH	220 pF
21.25 MHz	0.53 μH	0.52 μH	147 pF
28.80 MHz	0.39 μH	0.38 μH	108 pF

summary

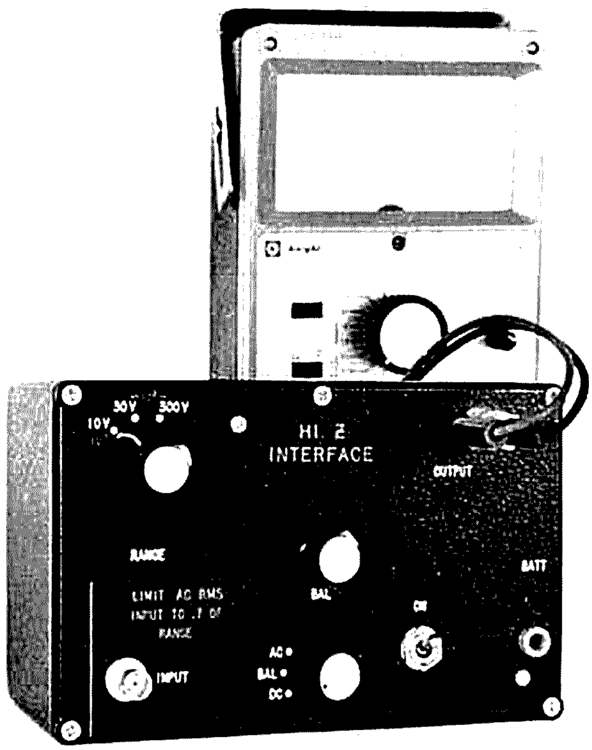
As indicated by the brownish lace pattern at the base of the tube envelope, the "well-used" 4-1000A broadcast tube I am using must have had a zillion hours on it when I first got it. It has been in use for over three years at my station and is still going strong. If it goes out tomorrow, I will have had more than my money's worth from it and the spare I purchased for $5 each.

Despite the work of building (and rebuilding) and debugging the modified 4-1000A grounded-grid linear it has been well worth the effort. My SB-401 exciter can drive the final to well over 2000 -watts PEP without strain, and can loaf along at half power and still drive the final to a full 1-kW input. When connected to my five-band antenna system on one tower[2] the results are quite gratifying.

references

1. I.L. McNally, "Graphical Solution of Impedance-Matching Problems," *ham radio*, December, 1969, page 26.
2. John R. True, W4OQ, "Grounded Vertical Tower Antenna System," *ham radio*, April, 1973, page 16; May, 1973, page 56.

ham radio



high-impedance meter interface

A high accuracy meter-interface unit for ac and dc measurements that features up to 1-million megohms input resistance

J.R. Laughlin, 11918 Pompano Lane, Houston, Texas

Making accurate measurements with a voltmeter is often difficult because of the input resistance of the meter itself. Errors due to the loading effect of the voltmeter are often nearly impossible to evaluate, particularly in transistor circuits where the dynamic characteristics of the circuit are difficult to accurately analyze. More importantly, many of these erroneous readings might go unnoticed as you hurry to check out a circuit. An accumulation of errors here and there often add up to a puzzling situation when troubleshooting a defective or inoperative circuit.

The standard vom with 20k V input resistance can be a disaster in many circuits, either requiring tedious calculations to correct for its loading, giving misleading results to those not completely familiar with the meter and circuitry or simply being useless as a measuring tool. Even the heralded vtvm, with its 10-megohm input resistance, will cause significant errors in many circuits.

Having long been troubled by voltmeter errors due to loading, I have made an effort to overcome and eliminate, as well as practically possible, this source of trouble. First, I decided that some type of interfacing unit should be designed to be used with existing equipment. Since my workshop contains a varied assortment of different types of meters, from the very cheapest to more expensive laboratory types, it would represent a needless waste of revenue to obsolete these. Also, the design of the interfacing unit would be greatly simplified by making use of the basic structure of these existing meters.

Portability, a must, dictates battery operation. To extend battery life and reduce operating costs, micropower operation is mandatory. The final instrument incorporates the basic specifications given in table 1.

circuit

To achieve extremely high input resistance a special dual fet with exceptionally low gate-leakage current was chosen to form the heart of this instrument (see **fig. 2**). For maximum linearity and accuracy the fet is combined with a high quality operational amplifier. The two are connected as a voltage follower with a gain of one. One important parameter of the op-amp is its very low power consumption. This results in truly low battery drain for the instrument as a whole, greatly prolonging battery life, reducing operating expense and eliminating the annoyance of frequent battery replacement.

The amplifier will handle an input voltage, without overload, of somewhat over 10 volts. When used within this range the input gate of the fet is connected directly to the circuit being measured. This mode of operation offers the highest input resistance obtainable from the amplifier. The only loading on the circuit being measured is the leakage current of the fet gate, this current being in the order of 0.2 picoamperes (0.0000002 microamps). Contrast this with the 50 microamps that a standard vom will draw, or the 0.1 microamps drawn by a vtvm when measuring one volt full scale.

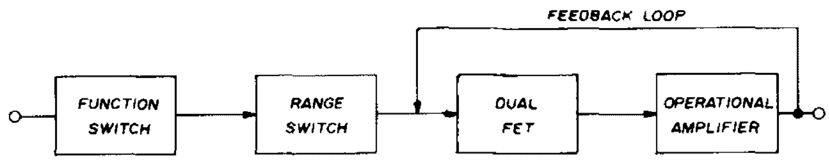
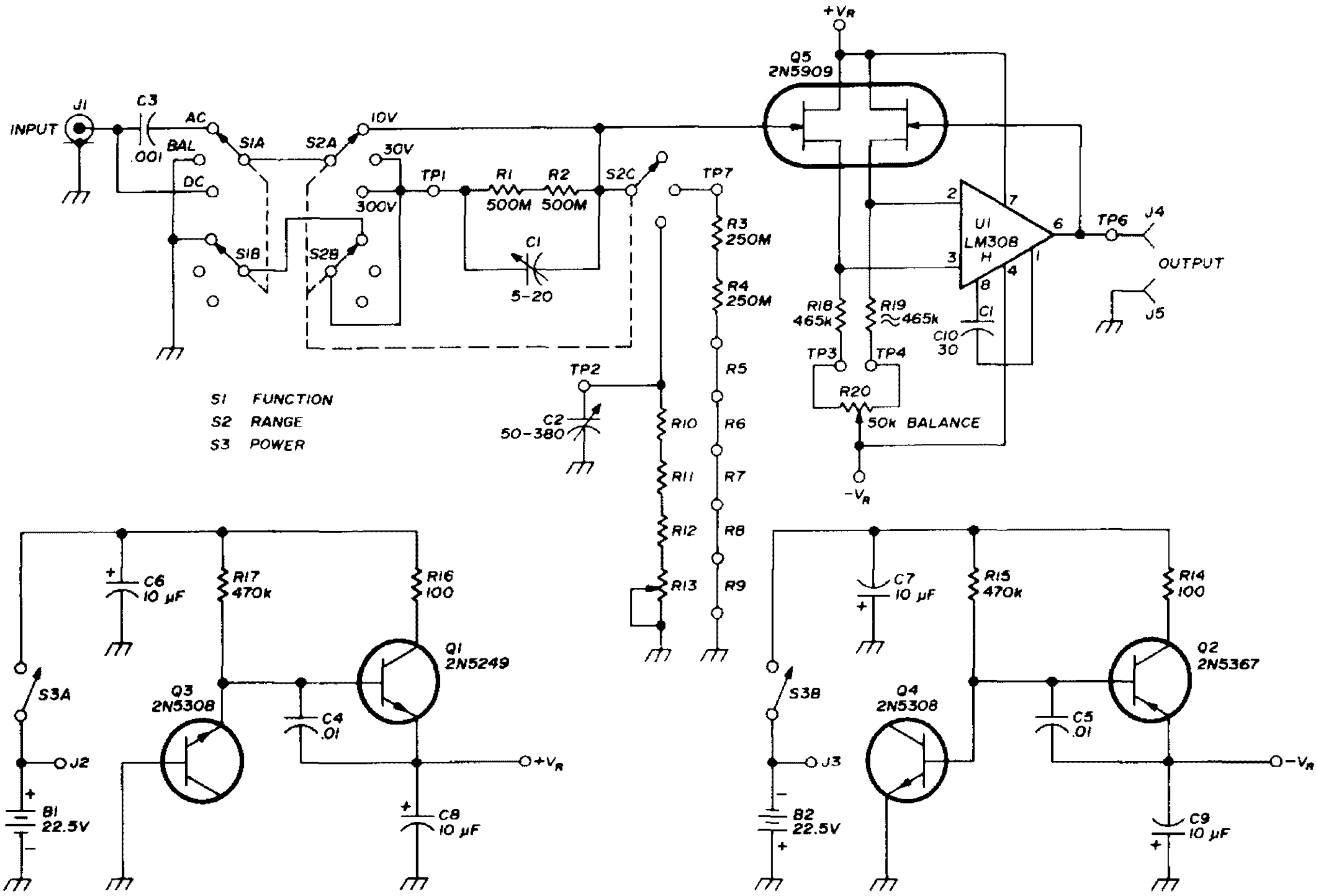


fig. 1. Block diagram of the high-impedance meter-interface unit. Instrument will solve dc measurement difficulties caused by circuit loading.

For measuring voltages in excess of ten volts a resistive voltage divider is necessary to keep the voltage level applied to the fet gate within its normal operating range. This voltage divider necessarily reduces the input resistance of the amplifier. To minimize loading on the higher ranges a voltage divider with approximately 1000 megohms total resistance was chosen. Unfortunately, this is considerably lower than the intrinsic input resistance of the fet, as enjoyed on the

table 1. Basic specifications of the high-impedance meter interface unit.

Dc input resistance, 0-10 volts	greater than 1-million megohms
Dc input resistance, 10 volts up	1000 megohms or greater
Supply voltage (battery)	±22.5 volts dc
Current drain	500 μA
Battery type	Eveready type 412
Battery life	295 hours, continuous usage
Accuracy, 0-10 volts	0.1% or better
Accuracy, higher ranges	depends on accuracy of dividers



C1	5-20 pF ceramic trimmer (Erie type 503). (Do not use a compression-type trimmer here.)
C2	50-380 pF compression trimmer (Elmenco type 465)
J1	BNC connector (Amphenol UG-657/U)
R1,R2	to total 1000 megohms, 1% (matched pair)
R3,R4	to total 500 megohms, 1% (matched pair)
R5-R9	values as required for balance (see text)
R10,R11,R12	to total 33 megohms, 5%
R13	2-3 megohm trimmer (Mallory 26L1 or 26L4)
R14,R16	100 ohm, 10% carbon
R15,R17	470k, 10% carbon
R18	465k, film-type preferable
R19	465k, approximately (see text)
R20	50k, 10-turn potentiometer
R21	220k, 10% carbon
S1	2-pole, 3-position ceramic wafer rotary switch
S2	3-pole, 3-position ceramic wafer rotary switch
S3	dpdt toggle switch

fig. 2. Schematic diagram of the high-impedance meter-interface unit. Three dc ranges are provided: 10, 30 and 300 volts; ac rms inputs are limited to 70% of dc range.

zero to 10-volt range, but it is still 100 times greater than the standard vtvm input resistance.

Although a higher value of resistance can be used in the divider network, this will cause degraded performance at higher operating temperatures. This brings up a point about fet gate leakage vs temperature – the gate leakage increases approximately 10 times for each 18°F increase in temperature. Hence, this instrument should be kept cool (room temperature) for maximum input resistance on the zero to 10-volt range.

Two voltage regulators are used in the battery supply circuit. These were not put there because of a great need for supply regulation, although the regulation will certainly tighten drift specs some, but the primary purpose of the regulators is to allow use of battery voltage in excess of the voltage rating of the op-amp. The

absolute maximum supply voltage rating of the LM308H op-amp is ±18 volts.

The closest available battery voltage that does not exceed this rating is a 15-volt unit. Use of a 15-volt supply would have placed the operating voltage of the amplifier uncomfortably close to the desired 10-volt input signal handling level. Of course, this 15-volt supply would have been the *starting* voltage of the battery and as the battery began to "droop" with usage the already uncomfortable margin would quickly become more narrow. A marginal situation like this would contribute to shortened battery service and danger of overload with normal input signal levels.

The input network consists of a resistive voltage divider with frequency compensating capacitors for better ac performance (frequency response), switching arrangement and blocking capacitor for ac amplification only, if desired. On the zero to 10-volt ac range the two 500-megohm resistors are switched from fet gate to ground. This provides the necessary dc path to ground for the fet gate, which otherwise would not exist due to the presence of the dc blocking capacitor, C3.

Resistors of the extremely high values used in this divider are not common stock items. Their procurement can be a real problem through standard channels. All of the resistors used here were supplied by the Resistance Products Company.* This organization specializes in very high resistance products and can supply them in almost any tolerance desired. The parts list describes the type and part number of the resistors I used.

The actual total resistance of the resistive divider is not as important as the fact that it must be high and that the ratio between R1 + R2 and the two ground legs must be close to that required for the proper division ratio. As used in this particular instrument, the voltage division ratios are 3:1 for the zero to 30-volt range and 30:1 for the zero to 300-volt range. This places the ratio between R1 + R2 and the two ground legs at 2:1 and 29:1. For a 1000-megohm value for R1 + R2 the zero to 30-volt ground leg is exactly 500 megohms. For the zero to 300-volt range the ground leg is 34.48 megohms.

*Resistance Products Company, 914 South 13th Street, Harrisburg, Pennsylvania 17104.

The high-impedance meter interface instrument is built into a cast aluminum box, Bud type CU347.

The following is a general formula for figuring the resistance ratios for any voltage ratio:

$$\text{Resistance ratio} = E_{in}/E_{out} - 1$$

$$\text{Ground leg resistance} = \frac{R1 + R2}{\frac{E_{in}}{E_{out}} - 1}$$

These simple expressions will allow easy computation of resistors required for any situation.

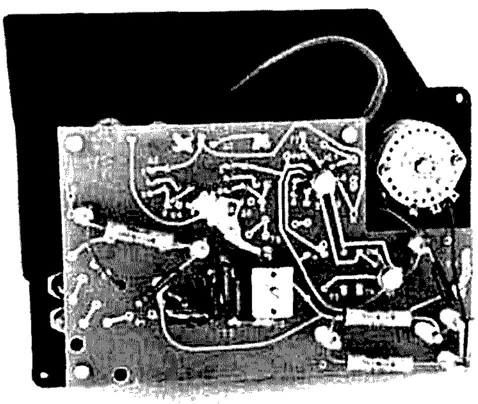
The instrument described here was designed to be used primarily with a readout meter having scales ending in 10 and 30. Hence, the ranges of 10, 30 and 300. For meter scales different than the above, the ranges of the interface will undoubtedly need to be scaled to match.

construction

The instrument is housed in a Bud Radio type CU347 cast-aluminum box. A good paint job with engraved legends will result in a professional appearance. The front panel must be drilled to accomodate the input jack, function switch, balance pot, range switch, power-on toggle switch, battery check jacks and out-

put terminals. Care should be exercised here to insure that the switches are not mounted too close to the edge so as to obstruct the case sides.

The circuit board is mounted on standoffs 1.5-inch long. This provides adequate clearance between the board and panel-mounted items. Location holes for the standoff legs fall in the center of the small round mold marks on the inside surface of the front panel. Also, the circuit board can be used to dimension these holes. High quality Teflon standoff terminals were pressed into the circuit board to hold R1, R2 and R3. Input wiring was run point-to-point instead of bundling to reduce interwiring capacitance and leakage currents. Teflon wire was used throughout. Be certain to use only ceramic switch wafers on the range and function switches. The input jack should have a Teflon insulator.



Most of the major components of the high-impedance interface unit are built on a printed-circuit board (see fig. 3). Switches, balance control and jacks are mounted on front panel.

Extra positions on the circuit board accommodate a number of resistors that may be connected in series to trim the exact value of the ground leg resistors in the voltage dividers. If quality resistors are purchased all of these positions will probably not be used.

Note. The circuit board contains pads for compensation components that are used on other types of micro power op-amps other than the LM308H shown in the parts list.

Here are some tips on placement of parts. The input gate is not connected to a board mounted pad but to a Teflon standoff. Room was provided on the board to accommodate very long resistors for R1 and R2. Some of these may be found surplus or purchased as replacements for elements in high-voltage probes. If short resistors are used as suggested in the parts list the standoffs will have to be positioned closer to the fet.

After mounting R1 and R2, C2 may be soldered directly to the standoffs holding these resistors as shown on the parts location diagram. When mounting the fet, Q5, be certain to form the leads so that they enter the proper pads on the board without shorting to each other. If the 2N5909 is used as suggested, the legs will not fall directly into the proper pads.

After completion of all wiring, the circuit board and all standoffs and switch wafers should be thoroughly cleaned with alcohol (pure) to remove all trace of rosin, fingerprints or other contaminants. The high megohm resistors should be cleaned also as their value can be significantly altered by contaminants on their surface.

checkout

First, measure battery drain imposed on each battery. This current should be approximately 0.5 mA. Significantly higher currents indicate trouble and should be investigated before proceeding.

The output voltage of the regulators should be measured. This voltage level will vary but will usually fall between 15 and 18 volts. If higher than 18 volts, the cause should be found and steps taken to being it to normal.

With the function switch in the *BAL* position, check to see if the balance control will vary the output + and − by approximately the same amount. Normal variation is around ±50 millivolts.

On the 10-volt range, dc input voltages both, + and −, should be reproduced exactly at the output up to the overload point. Overload will normally occur at about 1 or 2 volts below the supply

voltage. Millivolt differences between input and output can be adjusted to zero with the balance control.

The two higher ranges should be checked, and trimmed if necessary, for accurate voltage division ratios. This procedure calls for an accurate and linear readout standard.

Input resistance levels as encountered here are difficult to measure with standard equipment. A simple check to demonstrate the high input resistance is to connect a standard 10-megohm vtvm in series with a 10-megohm resistor and apply a voltage to the series combination. Note the voltage reading on the vtvm; it should be approximately half the supply voltage. Now, alternately connect and disconnect the high-impedance interface to the probe of the vtvm while noting the change in reading of the vtvm. The 1000-megohm input resistance of the two higher ranges should cause only about a 1% change in the reading of the vtvm. The input resistance of the 10-volt range should cause no noticeable variation in the vtvm reading.

An audio oscillator is needed for checking the ac characteristics of the interface. Stray capacitance plays an important role in the frequency response of the instrument. Consequently, the trimmer capacitors C1 and C2 must be adjusted a little at a time and the instrument put into its cabinet after each adjustment to note the effect. There is no adjustment necessary for the 10-volt range.

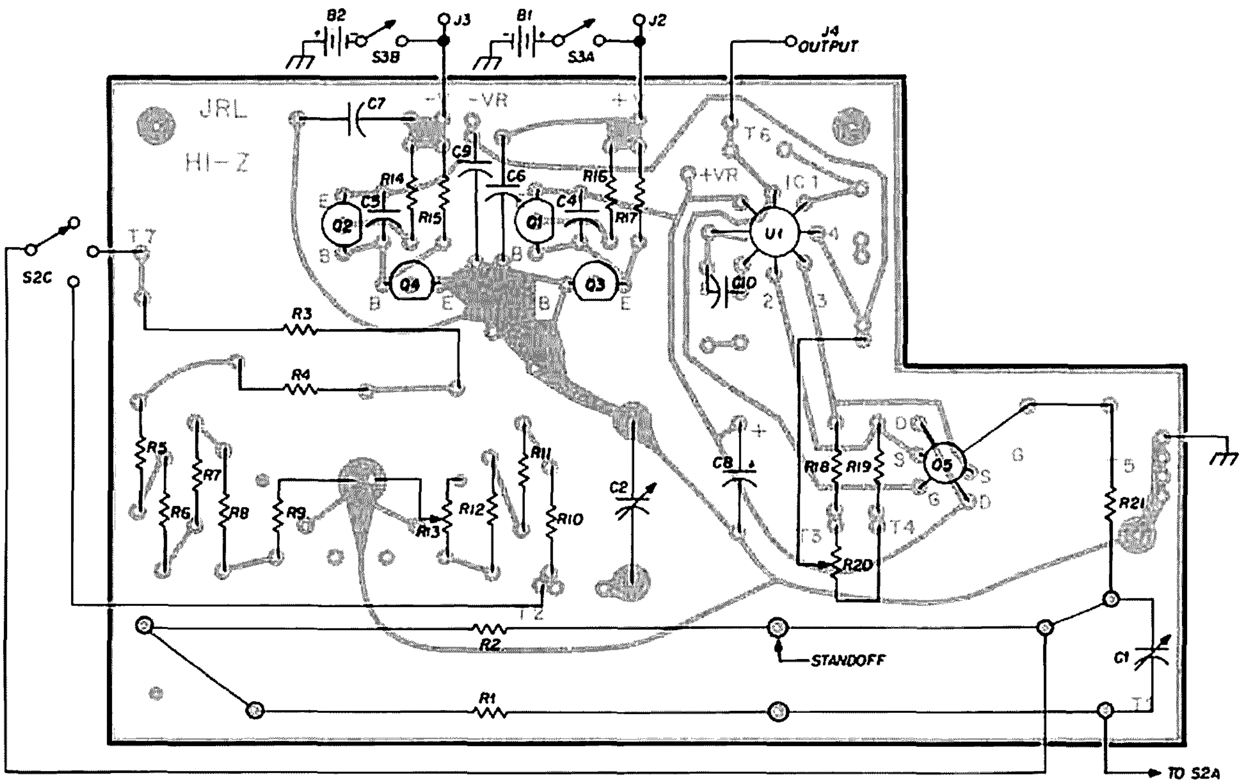


fig. 3. Printed-circuit layout for the high-impedance meter-interface unit.

The simplest method of adjustment can be accomplished by feeding a 100-Hz square wave into the instrument and adjusting C1 on the second range for best response with the instrument in the cabinet. Next, C2 is adjusted with the range switch on the highest range for best square wave response, with the instrument in the cabinet. Be certain that your oscilloscope has sufficient response to exactly reproduce the square wave as it comes out of the generator.

If only a sine wave oscillator or a 60-Hz source is available, the adjustments can be made so that exact reproduction of the input amplitude occurs at the output terminals with the instrument in the cabinet.

ham radio

IC logic families

A rundown on popular logic circuits, their interfaces, and compatibility

Hank Olson, W6GXN, Stanford Research Institute, Menlo Park, California 94025

The use of the Fairchild μL914, and other members of the resistor-transistor logic (RTL) family, is well documented in the amateur radio periodicals. The Fairchild μL900, μL914, and μL923 (all from the RTL family) were the first digital integrated circuits to be offered in low-cost epoxy packages. An early article by Lancaster[1] pointed the way for many other experimenters, and for years RTL was the most-used logic family among hams. A number of other firms rushed into production of RTL to "second-source" these popular ICs; all but several of them dropped out of the competition in a year or two. Motorola went Fairchild one better – offering an expanded RTL family in plastic dual-inline packages (DIP). Some of these plastic DIPs are also offered as part of the Motorola HEP line, which are more easily obtained than most other ICs. The HEP line also has RTL ICs in the TO5 metal can package.

The following selection guide should be helpful when planning your next project using ICs. A brief description is given of seven popular logic families together with information on interchangeability and compatibility. Additional data on device details and applications is available from the manufacturers mentioned.

RTL logic

Although RTL is widely used in ham circles, and also fairly inexpensive, it has some limitations. RTL is a relatively slow form of logic. While some members of the family are rated up to 8 MHz, the family is not usually used above 1 or 2 MHz – especially the low-power versions. RTL requires a supply voltage of 3.6 Vdc at relatively high current. A modest logic array of RTL devices can often require

several amperes. RTL has poor noise immunity, the level of ripple or transients on the power supply bus above which false triggering can occur. This low ripple limitation, together with the large current requirement, can make the power supply quite expensive.

The nice features of RTL, other than economy, are the ease of understanding and durability in the hands of the beginning electronic logician. As such, it is no wonder so many technical people cut their teeth on RTL. **Fig. 1** shows an RTL two-input gate; it is similar to one-half of a μL914 or HEP584. This circuit could also be built of discrete components – say a pair of 2N708s, a 470-ohm resistor, and two 620-ohm resistors (to use standard components).

Operation. If both inputs are grounded (or open circuited), no *current will* flow in either transistor, and the output voltage level will be at +3.6 V. Now, let input 1 be raised to +2.2 V. This will forward-bias Q1, the 450-ohm load resistor will conduct current, and the output voltage will drop. Assuming the base-to-emitter drop to be 0.6 V, that puts 1.6 V across the input resistor: base current = 1.6V/640 ohms = 2.5 mA. This amount of base current is more than enough to saturate Q1, and output will drop to nearly ground level, say to +0.5 V. If we now make input 2 rise to +2.2 V, little change in the output will occur. So, if input 1 *or* input 2 is high, the output will be low.

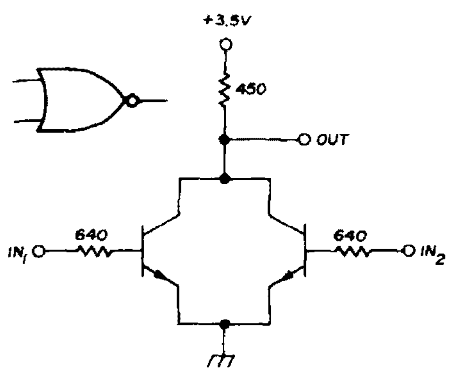


fig. 1. RTL 2-input NOR gate.

We will now define high as 1 (or true) and low as 0 (or false). Then, looking at our RTL gate we can say that if input 1 *or* 2 is 1, the output is 0. Since the output gives a false output to either of two true inputs we must call this gate a NOR gate (short for NOT-OR). Similarly, an inverter, as shown in **fig. 2**, is sometimes called a NOT gate. Such a NOT gate is contained in a Motorola MC789P, hex inverter (there are six in one package).

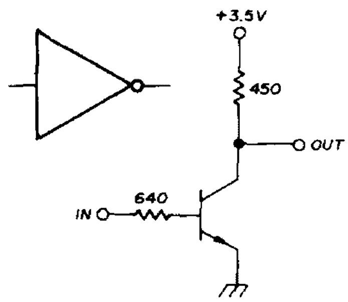


fig. 2. RTL inverter (NOT) gate.

By combining a NOR and NOT circuit we have an OR, as shown in **fig. 3A**. The output is 1 (true) if either input 1 *or* 2 is 1 (true). Similarly, by using a NOT gate ahead of each of the two inputs to the NOR gate, we can form an AND. This is shown in **fig. 3B**. The output is 1 (true) if both input 1 *and* input 2 are 1 (true).

A NOT gate can be made using a two-input NOR gate by simply grounding one of the two inputs. Therefore, all of the circuits thus far described can be constructed using one or more sections of μL914 gates. If one wanted just one NOT, NOR, or OR function, the μL914 would be the least expensive way of implementing it. Similarly, if just one AND function is desired, an MC724P is the cheapest way to build it. However, in larger systems (where a number of IC packages are used) gates are not usually used as inverters, since a hex inverter costs about the same as a quad two-input gate.

Although they are not *usually* drawn that way in logic diagrams, any of the several flip-flops can be made up of gates.

Fig. 4 shows how a number of flip-flops are made up, along with their usual logic circuit symbols.

Load factor. One of the nice features of IC logic is the simple system of fan-in and fan-out numbers that most manufacturers provide. In RTL, the Fairchild and Motorola fan-in and fan-out numbers are compatible. A fan-in of 1 is the loading that a low-power gate (μL910, say) puts on whatever is driving it. The fan-out is the number of such gates (as loads) an IC is capable of driving. For instance, the fan-out of a μL910 is 4; it can drive one input of another μL910 plus one input of a μL914 (fan-in of 3).

These load factors apply only within the RTL family, however, and more care must be exercised when trying to interface RTL with other logic families. This is not just an academic problem – say getting from an RTL system into a TTL system – because field requirements often place unlike pieces of apparatus together. Fortunately, there are a number of family-interface ICs available, and otherwise the transition circuitry can be built using discrete components.

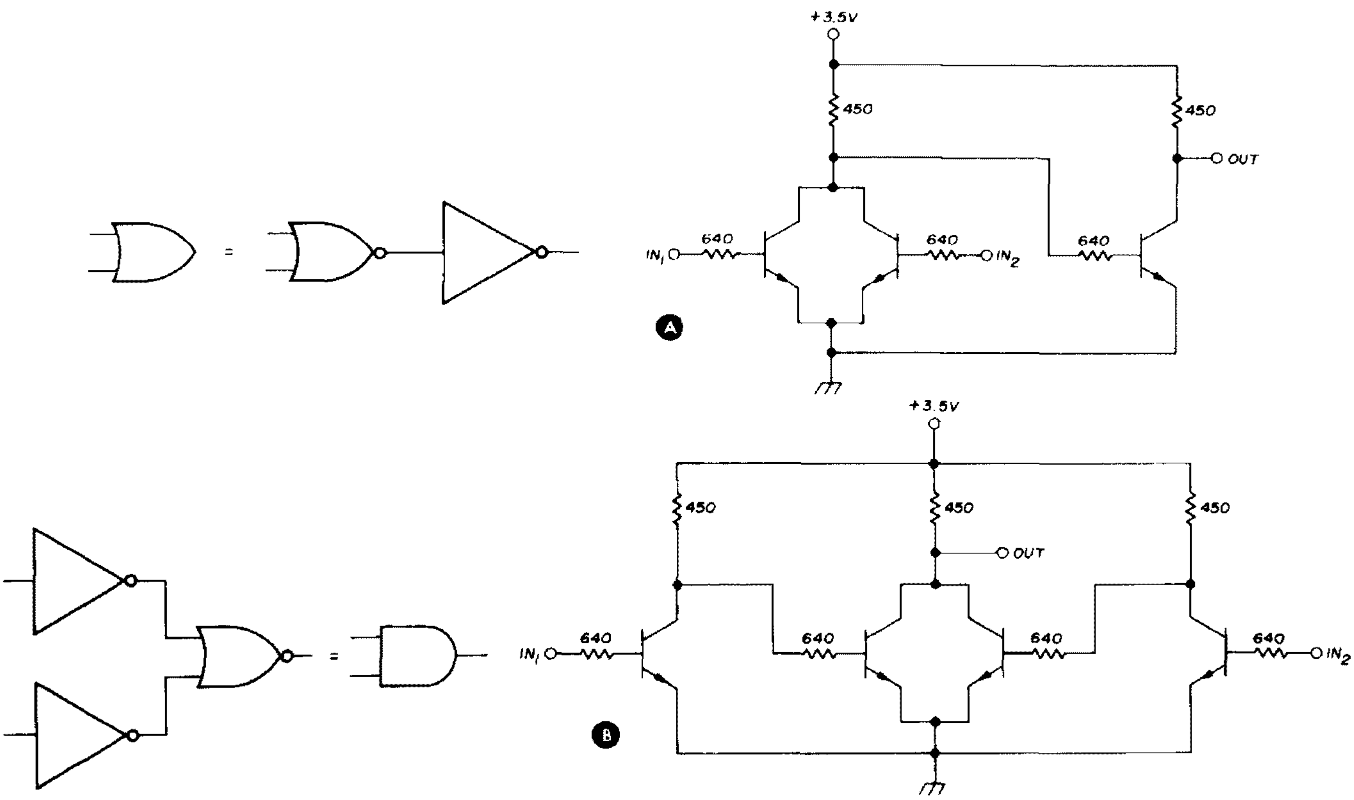


fig. 3. RTL OR gate (A) and AND gate (B).

DTL logic

Historically, diode-transistor logic (DTL) comes right behind RTL as an IC. However, DTL was extensively used as a logic form in discrete circuitry before ICs were developed. It is common to find old logic systems full of cards of 2N404s and 1N658s still in use in pieces of discrete DTL systems. Of course, all of the components on one whole card of such a system could be replaced today by one IC – less expensively! Such a discrete DTL gate is shown in **fig. 5.**

μL930 family. Usually, when one speaks of DTL, the μL930 family is meant. There are a number of other DTL families around, but they are the "losers" in the industry-wide acceptance game. Like the μL900 (RTL) series, Fairchild also originated the μL930 (DTL) family.

In the μL930 series, there are no round can packages available in plastic, although they are fairly common in

round metal can packages. The round metal can package is often referred to as TO5 because the can size is the same as that of a TO5 (three lead) transistor. While most round metal can ICs have 8, 10, or 12 leads and so cannot strictly be called TO5, the TO5 can description has been widely given to them. DTL is also available in the dual inline package (DIP) both in plastic and ceramic, and in the ceramic flat-pack.

Amateur DTL users will probably be most interested in the plastic DIP form of DTL, but often the other styles are offered as surplus at low prices. In an earlier article, I showed how to use surplus flat-pack μL930s and μL946s in an amateur wind-direction indicator.[2] Of course, the flat-pack style IC is not limited to DTL; nearly all logic families use it for their MIL-spec package.

Current-sinking logic. Fig. 6 shows a DTL gate (¼ of μL946 or MC846P). Note that both inputs are via diodes, which might at first look as though they are connected backward. Fear not, the diodes are shown correctly, because we must draw current out of the input when using this sort of logic. For this reason DTL is called current-sinking logic. (We will see that some other forms of logic are also current sinking as we go on.)

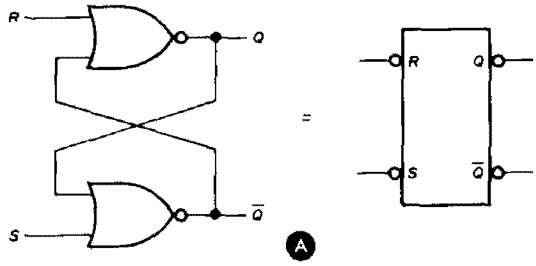
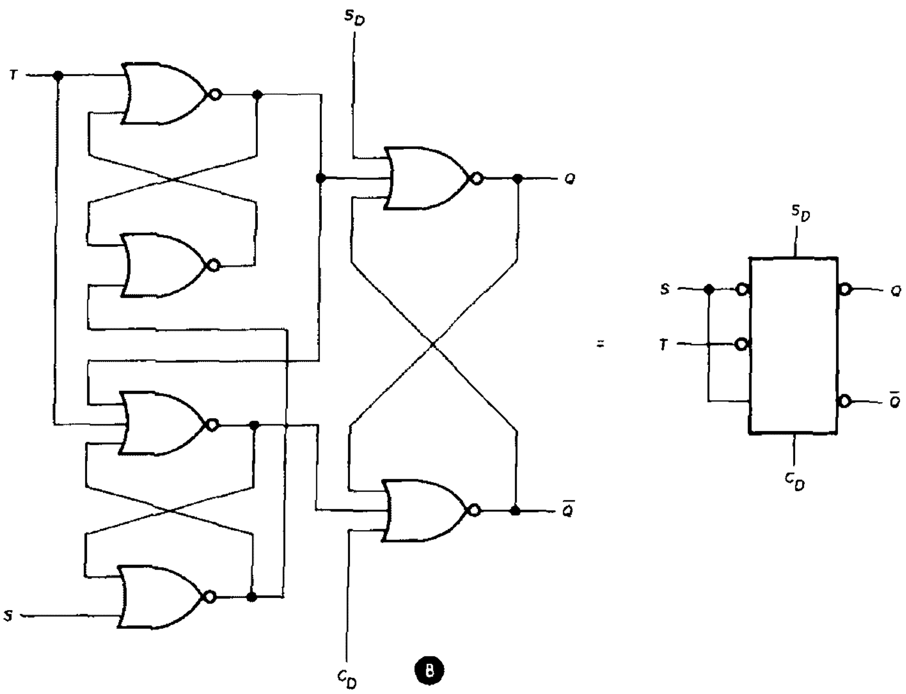


fig. 4. Flip-flop circuits. A set-reset (RS) type is shown in (A); a type D is shown in (B).



Unlike RTL, if a DTL gate has its inputs left open, the transistors in it are conducting, and the output is low or zero. Also, for positive logic, a simple DTL gate is called a NAND gate. When input 1 *and* input 2 are high (or open) the output is low. By following the gate of **fig. 6** with an inverter (such as 1/6 of an MC836P) we create an AND gate, as shown in **fig. 7.**

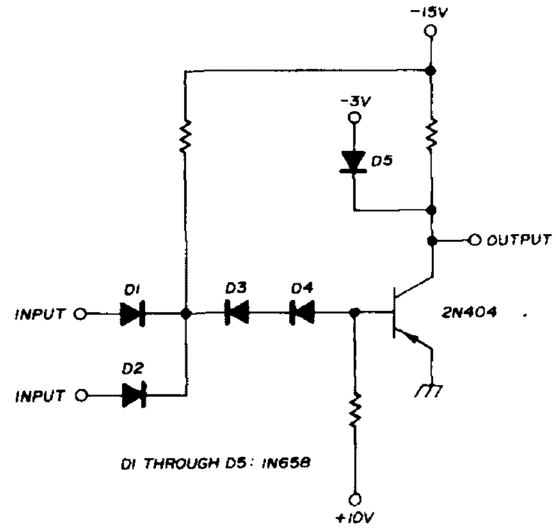


fig. 5. A DTL 2-input NAND gate using discrete components.

Similarly, by preceding each input of the NAND gate of an inverter, we create an OR gate, as in **fig. 8.** Note that this arrangement is just the opposite of the way we made the OR and AND functions with an RTL-NOR gate.

TTL logic

Transistor-transistor logic (TTL or T^2L) is a newer form of current-sinking logic. TTL was born in the IC age and has no discrete equivalent. A typical TTL two-input gate is shown in **fig. 9.** Note that the input transistor has two emitters. Each of these emitters acts much as one

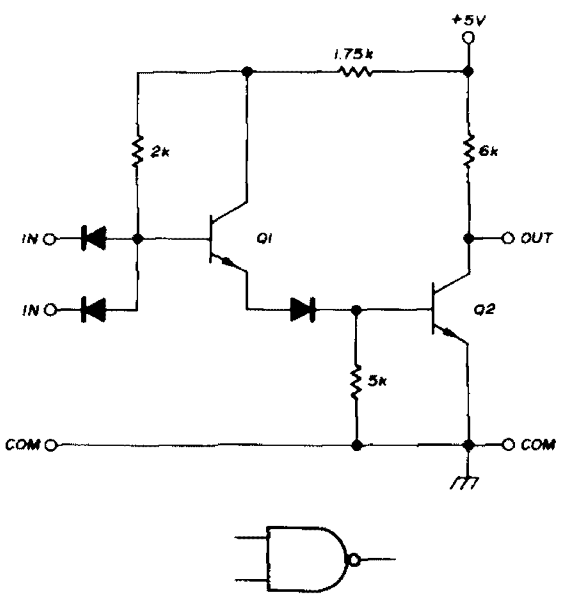


fig. 6. Example of current-sinking logic using ¼ of an MC846P DTL gate.

of the input diodes in a DTL gate to control the single input transistor.

Since TTL is also a current-sinking logic, and also operates on V_{cc} = +5 V, the logic levels are compatible with DTL. In general, TTL is faster than DTL, which imposes some restriction on the mixing of the two families, but at lower speeds they are compatible.

There are five major TTL families on the market today: Texas Instruments SN7400N, Sylvania SUHL, Fairchild TTμL9000, Motorola MC3000-4000, and Signetics DCL.

SN7400N and variations. The SN7400N line was originally produced by Texas Instruments, but most of this line is second-sourced by other firms. There are about 200 members of the SN7400N series, and many more variations. Varia-

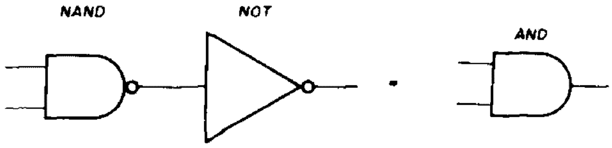


fig. 7. Formation of an AND gate using a NAND gate and an inverter.

tions are SN7400, SN5400N, and SN5400. The SN7400N is a plastic dual-inline package, SN7400 is a ceramic flat-pack, SN5400N is a plastic dual-inline package with MIL temperature specs, and SN5400 is a ceramic flat-pack with MIL spec temperature ratings.

In addition to the plastic dual-inline package, if one changes the N to a J at the end of the device number, the dual-inline package becomes ceramic. Further, if one puts an L between the SN74 and the next 2 or 3 digits, one gate is a lower-power version. If one puts an H in that same position, he gets a high speed version; and if an S is inserted, the ultra-high-speed Schottky version is specified that will toggle to 100 MHz. This method of variation is quite logical and creates a large variety from which to choose.

Design considerations. So that the electronic designer can "shift gears" when

adapting older DTL systems to TTL, Texas Instruments has devised a dual nomenclature system. Thus a quad 2-input gate is called both SN7400N and SN74-846N, so that the designer knows it has pin-for-pin compatibility (in addition to logic-level compatibility), which enables him to substitute SN7400N logic for μL930 logic on an etched circuit board with no changes (in many cases). Further, since TI also makes the μL930 DTL line, a family-to-family fan-in, fan-out number system is available.

In general, DTL can drive 8 DTL loads or 5 TTL loads; TTL can drive 10 TTL loads or 10 DTL loads. The DTL equivalent of the SN7400N, SN74-846N is (sensibly) the SN15-846N, for instance, if one is specifying TI parts. The L, H, and S versions of the SN7400N family (low power, high speed, and Schottky-clamped, respectively) are not as numerous as the standard series units, nor are they widely second-sourced.

In general, all the TTL families are compatible with each other (and with DTL). Details and pin arrangement are the biggest differences, but then details must be observed even when one is designing within one logic family.

HTL logic

A special variety of DTL used with higher supply voltage than the μL930

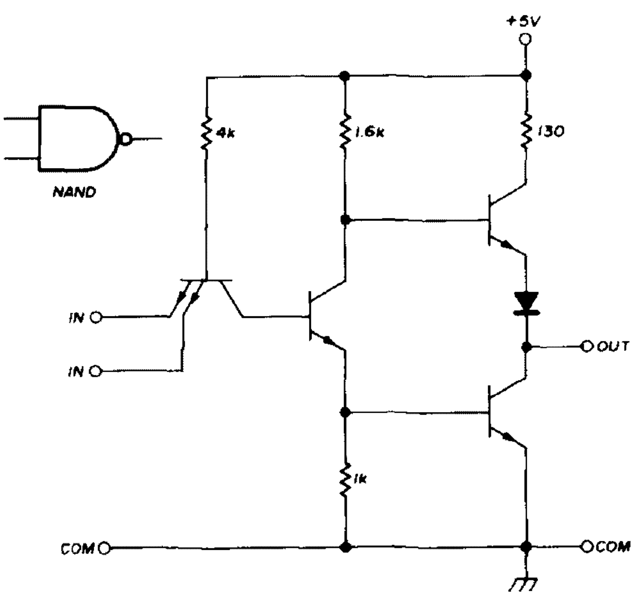


fig. 9. Typical 2-input TTL NAND gate consisting of ¼ of an SN7400N.

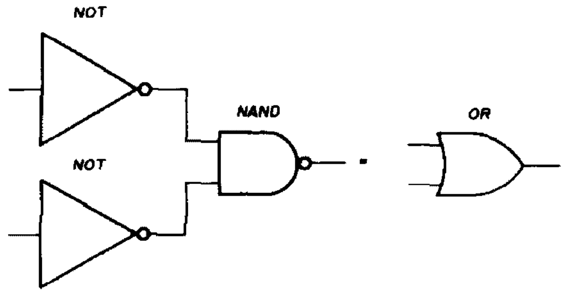


fig. 8. NAND gate inputs preceded by inverters equals an OR gate.

DTL is HTL (high-threshold logic). This family is, strictly speaking, offered only by Motorola, but HNIL (high-noise immunity logic) by Amelco Teledyne is essentially the same, as is HLLDTμL (high logic level diode-transistor micrologic) by Fairchild. These families all have the standard multiple diode inputs of DTL as shown in **fig. 10.**

Comparing **fig. 10** with **fig. 6**, we see the really important difference between these two DTL forms: the diode coupling Q1 and Q2 in **fig. 6** is replaced by a zener in **fig. 10.** This causes the HTL to have a much larger noise immunity. For this reason, HTL is often used in systems that contain electrically noisy components like relays and brush-type motors.

If we stick to V_{cc} of 12 to 15 volts, all these high-level families are compatible. The HNIL line will not operate above +15 V, however, as will the Motorola and Fairchild versions. All these HTL families have interface units available. Amelco and Motorola level-shift units will interface (either way) with DTL/TTL or RTL. The Fairchild level-shift units will interface only with DTL/TTL (either way).

ECL logic

Emitter-coupled logic (ECL) is the only bipolar IC logic form that does not use transistors that flip between the nonconducting and saturated states. This nonsaturated logic, also referred to as current mode logic, was pioneered by Motorola. Motorola has produced four series of MECL (Motorola emitter-coupled logic): MECL I, MECL II, MECL III, and MECL 10,000. MECL I operated at speeds up to 30 MHz, MECL II

operated at speeds up to 120 MHz, MECL III operates up to 350 MHz, and MECL 10,000 has lower power consumption but speed similar to MECL II and III. MECL I and II are second-sourced by Stewart Warner in pin-for-pin equivalents. RCA,

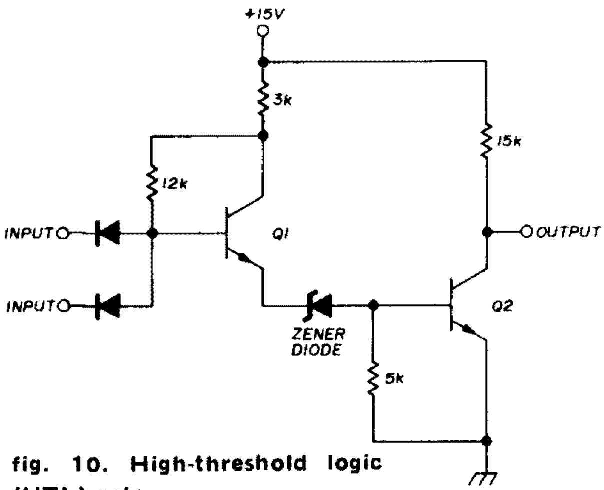


fig. 10. High-threshold logic (HTL) gate.

Fairchild, and Texas Instruments also offer their own lines of ECL, which have logic levels compatible with the Motorola family.

The curious thing about ECL is that the V_{cc} lead is usually grounded and the V_{ee} lead operated at -5.2 volts, yet ECL uses positive logic. Having the ECL units operating below ground makes ECL and other logic families awkward to operate together, because of the requirements for both positive and negative power supplies.

There are several logic-level translators available to interface ECL and DTL/TTL. MECL I has the single translators MC317 and MC318; MECL II has the single-level translators MC1017 and MC1018 plus the quad translator (MECL to DTL/TTL only) MC1039. There are no MECL III translators because there are no saturated IC logic families that can operate at speeds above about 100 MHz.

An ECL gate is shown in **fig. 11**; note R_e, the emitter resistor, from which the logic derives its name. The basic ECL gate is a differential amplifier, with the base of Q2 referenced to -1.15 volts. Since this differential pair is used in logic applications, either Q1 or Q2 is on. Since ECL uses positive logic (as do the other logic families we've already looked at), 1 = -0.075 V and 0 = -1.55 V. If either Q1 or Q1' are in the 0 condition, then RC2 draws all the current. In either case, the total current drawn by the pair is nearly constant.

Since there are two outputs in opposite states, this form of logic gate offers both NOR and OR functions as a basic part of its circuit. This feature is quite useful and can save the addition of extra inverters.

As seen in **fig. 11**, each output has an emitter follower built in to provide better fan-out. The typical fan-out of ECL is 15, which is considerably larger than most other logic families. Because fan-out from standard ECL units is so large, no buffer gates are offered. **Fig. 12** shows a Motorola MC306G in combination with a Motorola MC304G bias driver. The bias driver is simply a source of regulated reference voltage.

In the MECL II, III, and 10,000 lines, the bias driver is built into the gate chip, so no external source of V_{bb} is needed. Also, since the bias driver is built in, and the units are available in a plastic DIP, MECL II is generally less expensive than MECL I.

MOS logic

Another category of IC logic is the metal-oxide semiconductor (MOS). I avoid calling MOS a family, because no one company's MOS catalog has an industry-wide acceptance. The result of the absence of a dominant MOS logic family is chaotic – every company has its own particular idea as to what process makes the best chips.

There are some general things that can be said about the various MOS ICs, however. Most of them operate on a negative V_{dd} of -10 V to -40 V, therefore having negative logic (where 0 = 0 volts and 1 = a more negative voltage). They are generally slower than bipolar ICs and are well adapted to large-scale integration (LSI) because of the ease of getting a large number of MOS circuits per unit area on the chip.

Because MOS is so well adapted to LSI, wherein we are talking of hundreds or thousands of gate sections per chip, and because no dominant MOS family has emerged, we find that MOS is mostly available as large arrays, such as memories. A few companies, like Fairchild, offer individual logic circuits such as 3-input gates, 5-input gates, and dual flip-flops as building blocks. That is, these building blocks are offered specifically for the engineer who is mocking up a breadboard for a custom-made array.

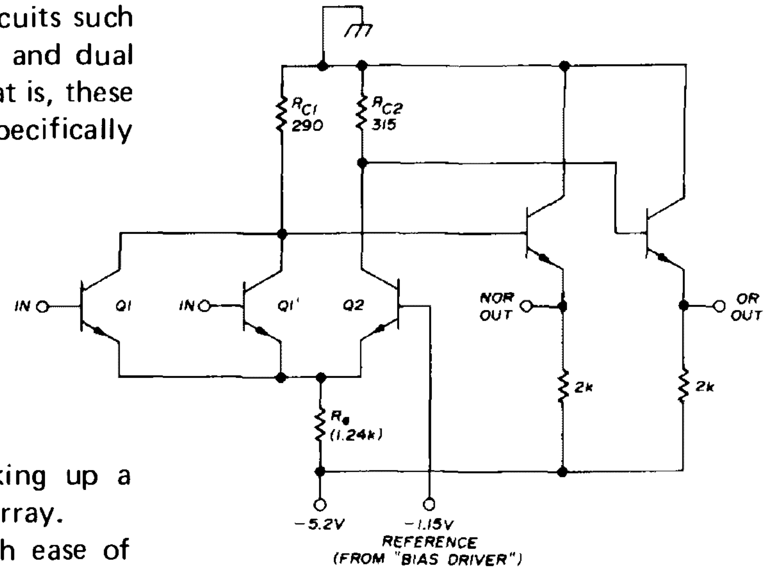


fig. 11. Emitter-coupled logic (ECL) gate.

One reason MOS offers such ease of LSI construction is that all the elements on the chip are tiny fets. Some are used as resistors and some as fets, but all are similar types. As an example, **fig. 13** shows a Fairchild 3102 three-input gate and how it can be used as a NOR gate, NAND gate, or how two units can form an RST (set-reset) flip-flop. The cost of the 3102 is $6.00 compared with an SN7410N triple 3-input gate at $0.60 (TI-TTL). It's no wonder MOS isn't used for small logic systems, since the cost per gate is quite high. However, in an LSI array the cost per gate can drop to pennies in production quantities.

CMOS logic

A new type of MOS logic that's gaining industry-wide acceptance is complementary MOS (CMOS). While conventional MOS logic owes its existence to the fact that many gates can be placed on one small chip, CMOS doesn't offer this advantage to as large a degree. The main selling point of CMOS is that it offers nearly zero standby power, the power drawn when the logic is in the 1 or 0 state. CMOS essentially draws current only during a transition from a 1 to a 0 or vice versa.

So, for logic systems that require low power, especially when fairly low data rates are to be handled, CMOS is very attractive. RCA was first to offer CMOS units, but now Motorola and others are also in the business, and the prices of these very fine logic ICs are likely to approach those of TTL as volume use develops.

CMOS operates on +5 to +15 V, as opposed to MOS, which operates on generally larger negative voltages. The +5 to +15 V supply voltage means that it's a simple task to interface CMOS with DTL, TTL, or HTL.

Fig. 14 shows a CMOS gate; note that it has in it both P-channel and N-channel mosfets. Like conventional MOS logic, the mosfets in CMOS logic are enhancement-mode types. This means they are like zero-bias triodes; they are nonconducting until the gate (grid) is forward biased. Enhancement-mode N-channel mosfets have been almost nonexistent as

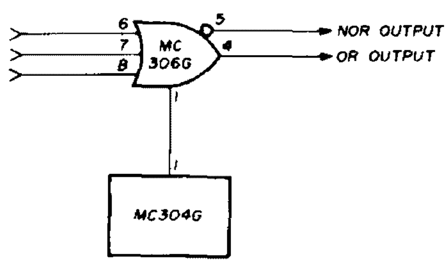


fig. 12. The MC304G bias driver provides a regulated reference voltage source for an MECL I gate.

discrete devices, which is probably why CMOS has become available only recently – i.e., a whole new MOS technology has had to be developed.

conclusion

We have looked at a number of logic families I feel are either common enough or inexpensive enough that amateurs will be likely to use them. There certainly are other logic families, which have been left out, because my intent was to present only the mainstream of IC logic; not the rare, expensive, or unique.

One often perplexing problem remains: that of interfacing logic families. Some comments on this have already been made in reference to ICs specifically made for family interfacing. The interfacing of RTL with DTL or TTL is fairly easy. My own preference is to use a μL900 (RTL buffer) to drive DTL or TTL from RTL. Driving RTL from DTL or TTL is usually direct; most DTL or TTL gates will provide enough voltage output to drive an RTL IC.

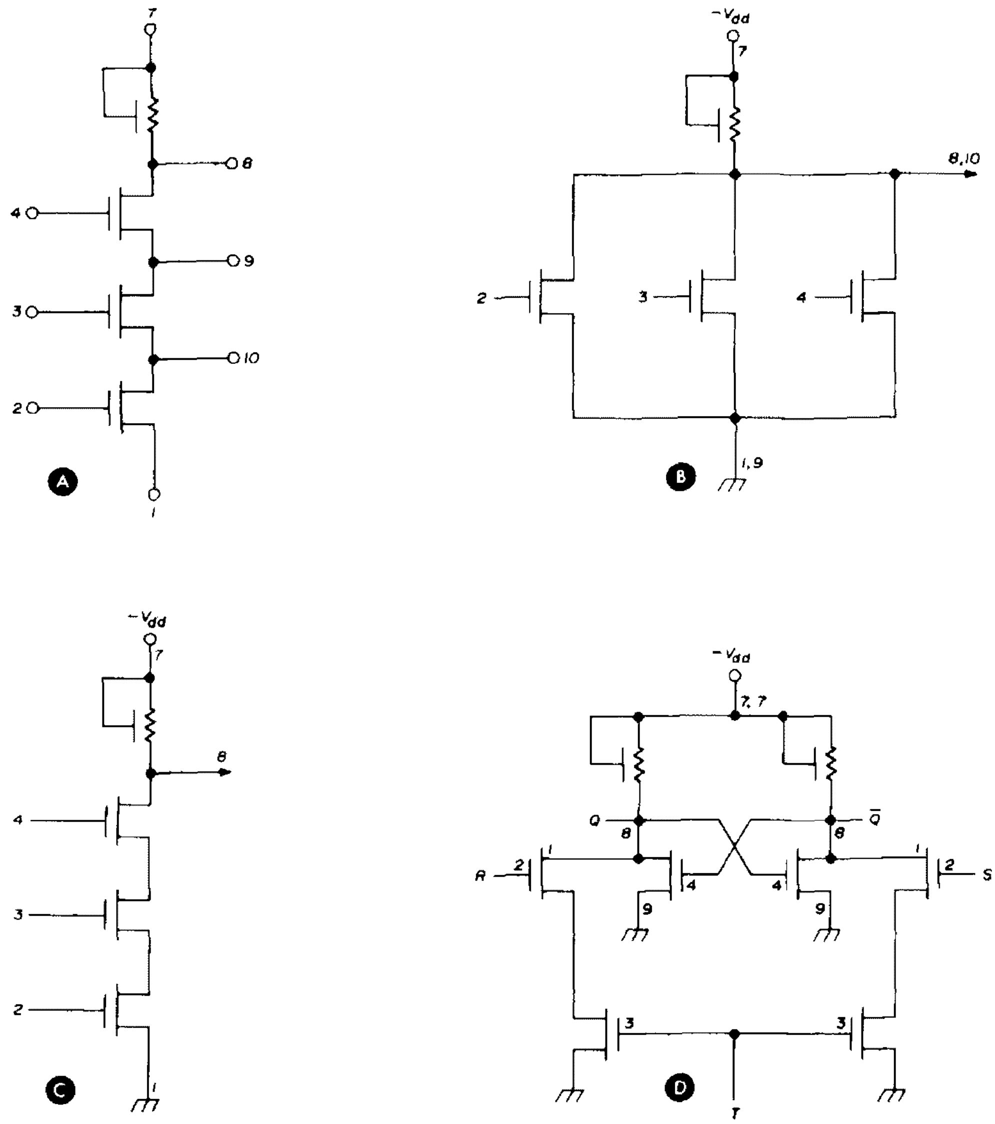


fig. 13. Fairchild 3102 3-input gate (A). A NOR gate, NAND gate, and RST flip-flop are shown in (B), (C) and (D).

Interfacing ECL and DTL/TTL has been mentioned in the discussion or ECL. As expected, the same interface ICs that drive DTL/TTL from ECL are also useful to drive RTL from ECL. The usual case is not to drive ECL from the lower-speed logic families (RTL, DTL, and TTL). The units that drive DTL/TTL and RTL from

ECL are MC317, MC1018, and MC1039. An example of the use of the MC1018 to convert between ECL and RTL is shown in reference 3.

There is even a readily available way to interface ECL and HTL. Motorola outlined this method at the 1969 series of digital integrated circuit seminars, using the MC1580. The method is shown in **fig. 15.**

The directly available DTL/TTL to HTL and RTL to HTL interface ICs have been mentioned in the discussion on HTL. These interface ICs can also be useful to drive some types of digital readout devices from RTL, DTL, or TTL.[4]

As pointed out in the discussion on MOS, there is no dominant MOS family yet; and so the interfacing of MOS with other logic forms is equally obscure – if not more so! In most cases the so-called bipolar-to-MOS interface circuits concentrate on converting the magnitude of shift from 0 to 1 and ignore the absolute values. This usually means that one ends up with the V_{ss} of the MOS portion of

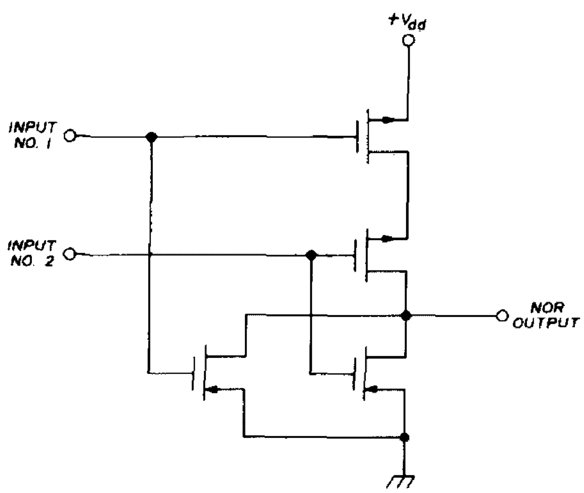


fig. 14. Complementary metal oxide semiconductor device (CMOS) 2-input NOR gate.

the circuit operating at +5 volts (when interfacing MOS and TTL) instead of at ground.

The interface may be constructed of discrete components as in reference **5.** Several companies offer hybrid ICs that are quite similar to discrete-wired interface circuits, which are quite expensive compared to monolithic ICs.[6,7] Fairchild has a pair of ICs available that convert nicely from TTL/DTL to MOS or MOS to TTL/DTL; these are respectively 9624 and 9625. Texas Instruments also makes a rather versatile IC called the SN75450,

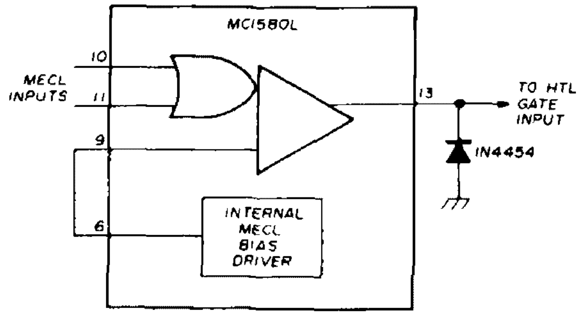


fig. 15. Interfacing MECL and HTL logic using an MC1580L.

which will interface MOS with TTL or vice versa. It requires a few external components, but it is very flexible.[8]

We've covered RTL, DTL, TTL, ECL, HTL, MOS, and CMOS families of IC logic in a sort of whirlwind fashion. Many logic ICs may also be used in circuits that are, strictly speaking, not logic functions, such as crystal oscillators. Many of the companies point up these uses in their data sheets, application notes, and handbooks.

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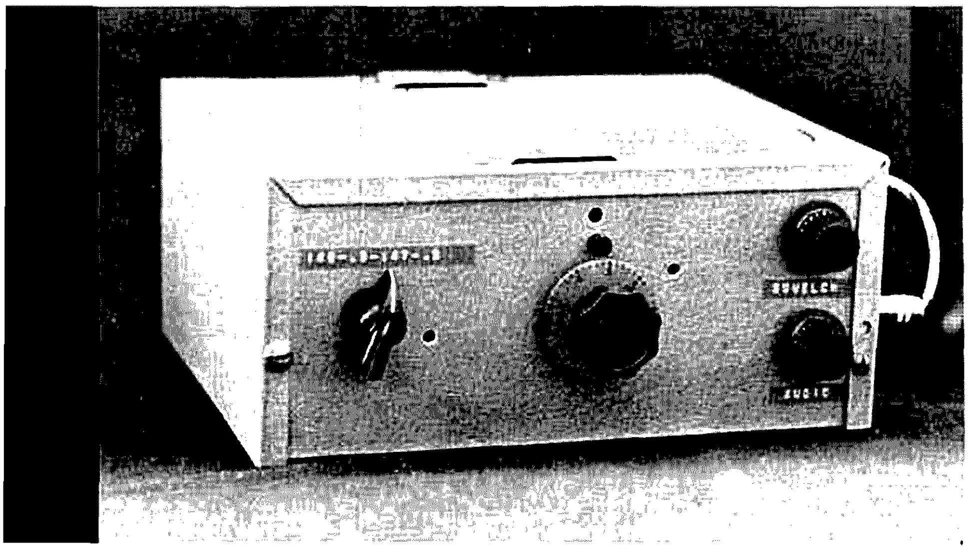


compact package

for two-meter fm

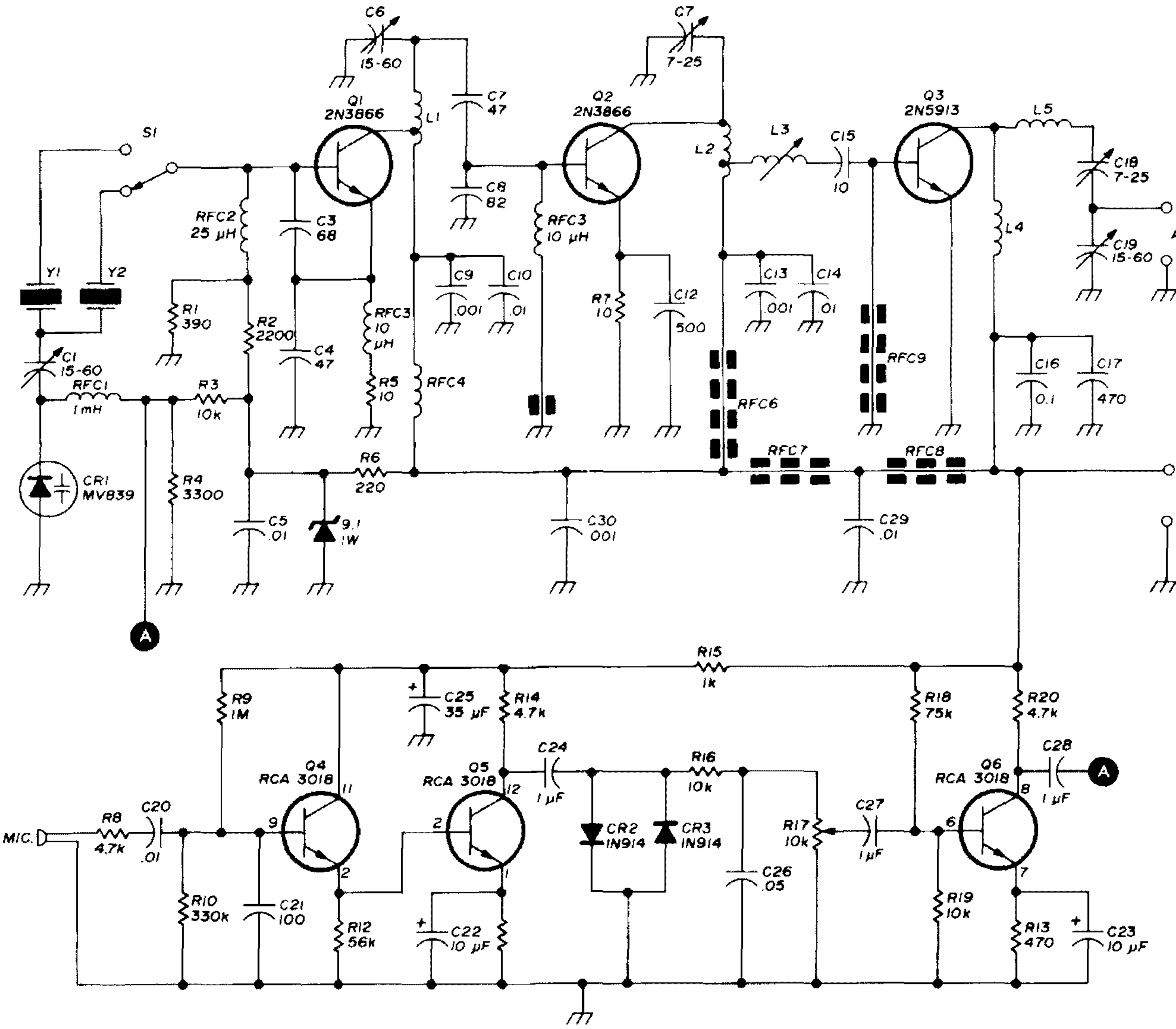
Design and construction details for a low-power fm transceiver for 144 MHz

J.H. Ellison, W6AOI, Menlo Park, California 94025

This article describes a compact package containing a low-power two-meter fm transmitter and a double-conversion, continuous-tuning fm receiver suitable for either fixed or mobile use. Although this description is primarily directed at the homebrew builder who balks at the cost of commercial equipment but lacks the time to develop his own design, the unit also includes several features which can be worked into different designs. The entire package, including everything but the power supply, is housed in a cabinet which is small enough to be accommodated in any automobile.

transmitter

Because this project is a combination of a previously described transmitter and an original receiver, design of the transmitter will be discussed first, and only with regard to certain minor changes and its performance. The transmitter is the Pip-Squeak, Mk II, described by W1CER.[1] The complete schematic is shown in fig. 1 for reference to the



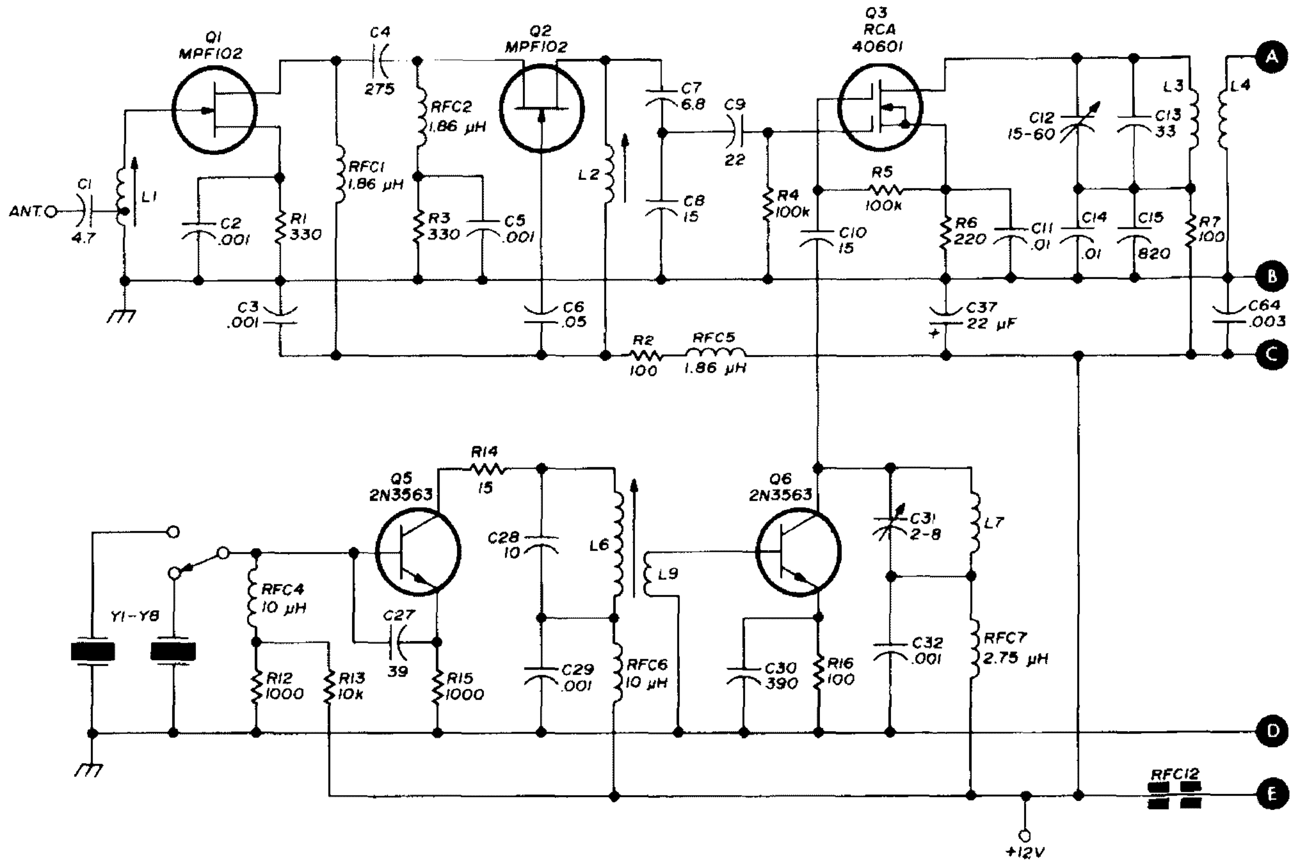
L1	5 turns no. 16 tinned bus wire, ¼" ID, 5/8" long, tapped at 2½ turns	L5	5 turns no. 16 tinned bus wire, 5/16" ID, ½" long
L2	3 turns no. 16 tinned bus wire, ¼" ID, 3/8" long, tapped ½ turn from C13 end	RFC6 RFC9	4 Amidon ferrite beads on ½" length of no. 24 wire
L3	4 turns no. 22 enamelled wire, closewound, ¼" ID	RFC7 RFC8	3 Amidon ferrite beads on ½" length of no. 24 wire
L4	25 turns no. 28, closewound on body of 100k, 1-watt resistor	Y1,Y2	18-MHz crystal ground for 20-pF load capacitance

fig. 1. Schematic for the two-meter fm transmitter. Transistors Q4, Q5 and Q6 are part of an RCA CA3018 IC.

changes mentioned above. In this circuit transistors Q4, Q5 and Q6 are part of an RCA CA3018 IC. Two 1N914 diodes should be used for symmetrical clipping rather than using the emitter-base junctions of transistors within the CA3018 IC.

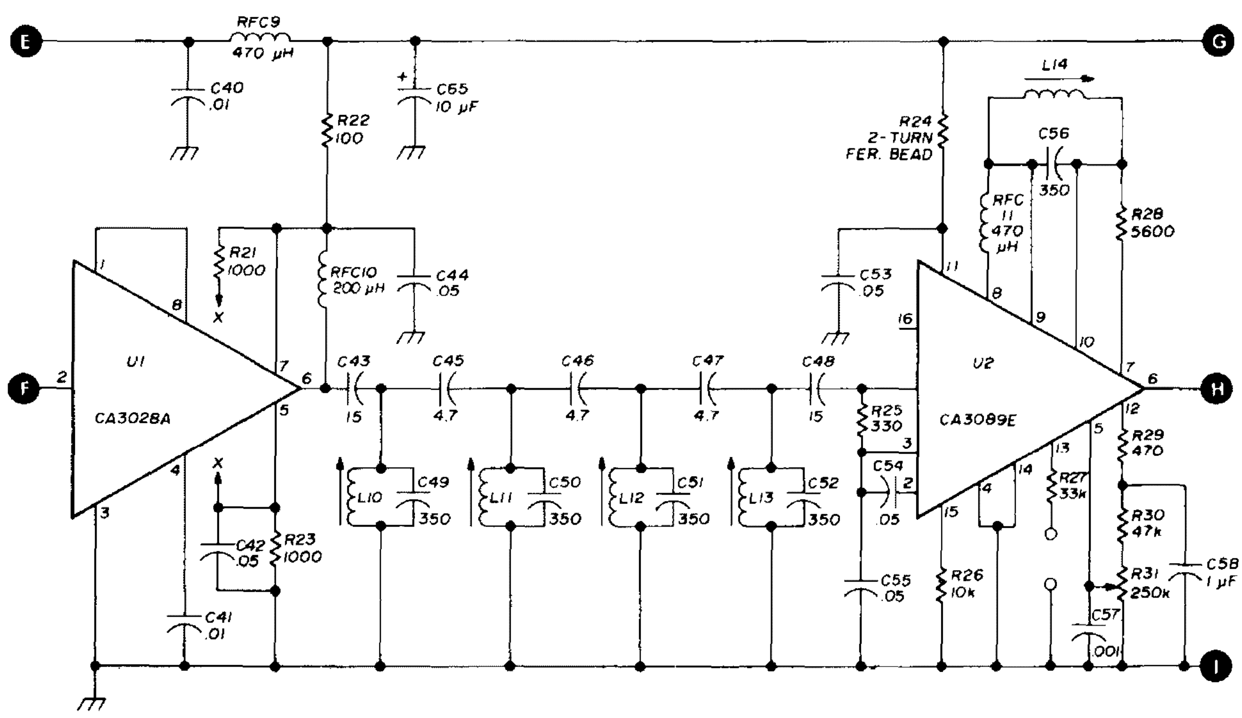
When building the unit, RFC5 and RFC9 should be placed on the foil side of the PC board. Emitter resistor R7 should consist of two ¼-watt resistors in parallel, dressed flat against the board for best stability. For further construction details, refer to the original article.

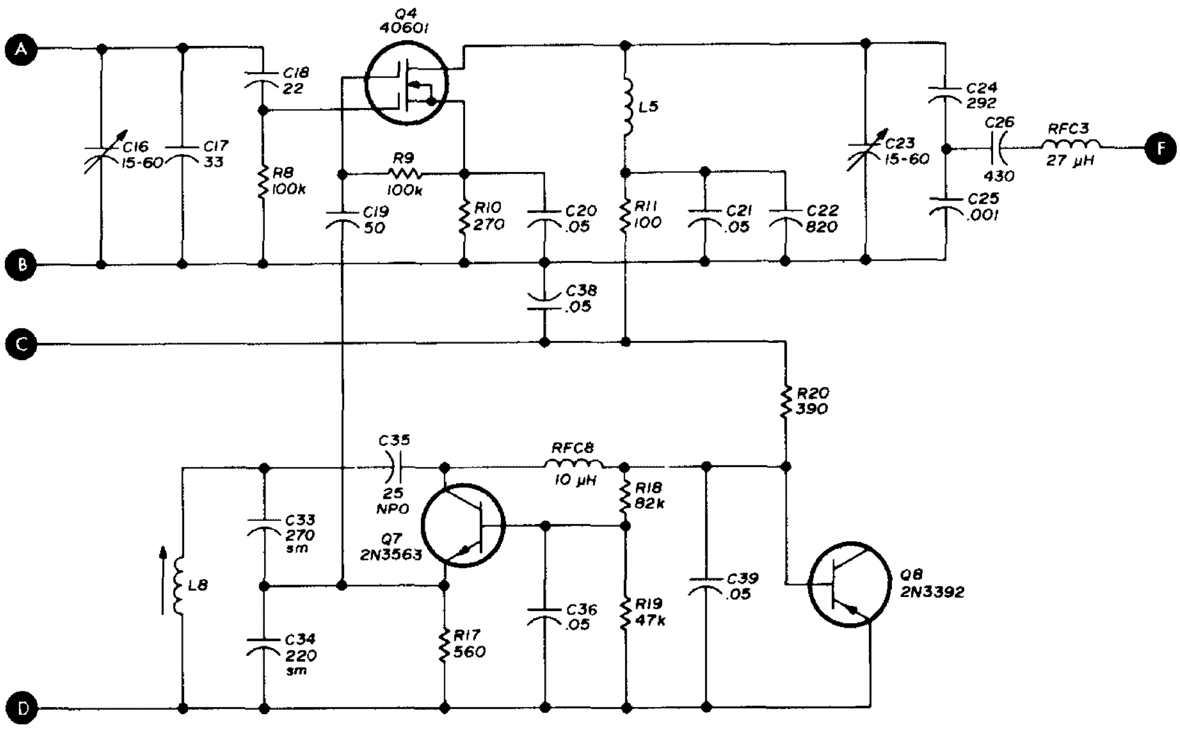
With these simple circuit changes, performance was easily duplicated in another transmitter which I built for a friend. The power output with a 12-volt power sup-



L1,L2	5 turns no. 20, 1/4" ID, 1/2" long	L8	14 turns no. 32 closewound on 13/32" slug-tuned ceramic form (Millen 69041)
L3,L4	23 turns no. 30, closewound on 0.219" ceramic form (no slug), 1/4" between windings	L9	2 turns no. 30 wound over cold end of L6
L5	38 turns no. 30 wound on 0.375" toroid core	L10,L11, L12,L13, L14	28 to 60 μH (J.W. Miller 9054)
L6	11 turns no. 28, closewound on 0.219" ceramic form		
L7	6 turns no. 20, 1/4" ID, 1/2" long	RFC12	2 ferrite beads

fig. 2. Tunable, continuous-coverage receiver for two-meter fm.

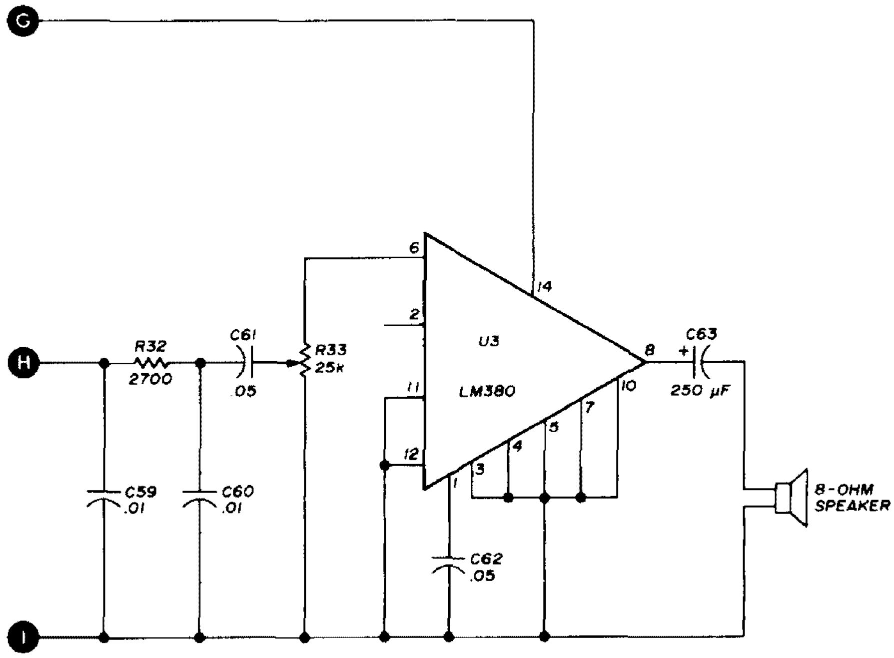


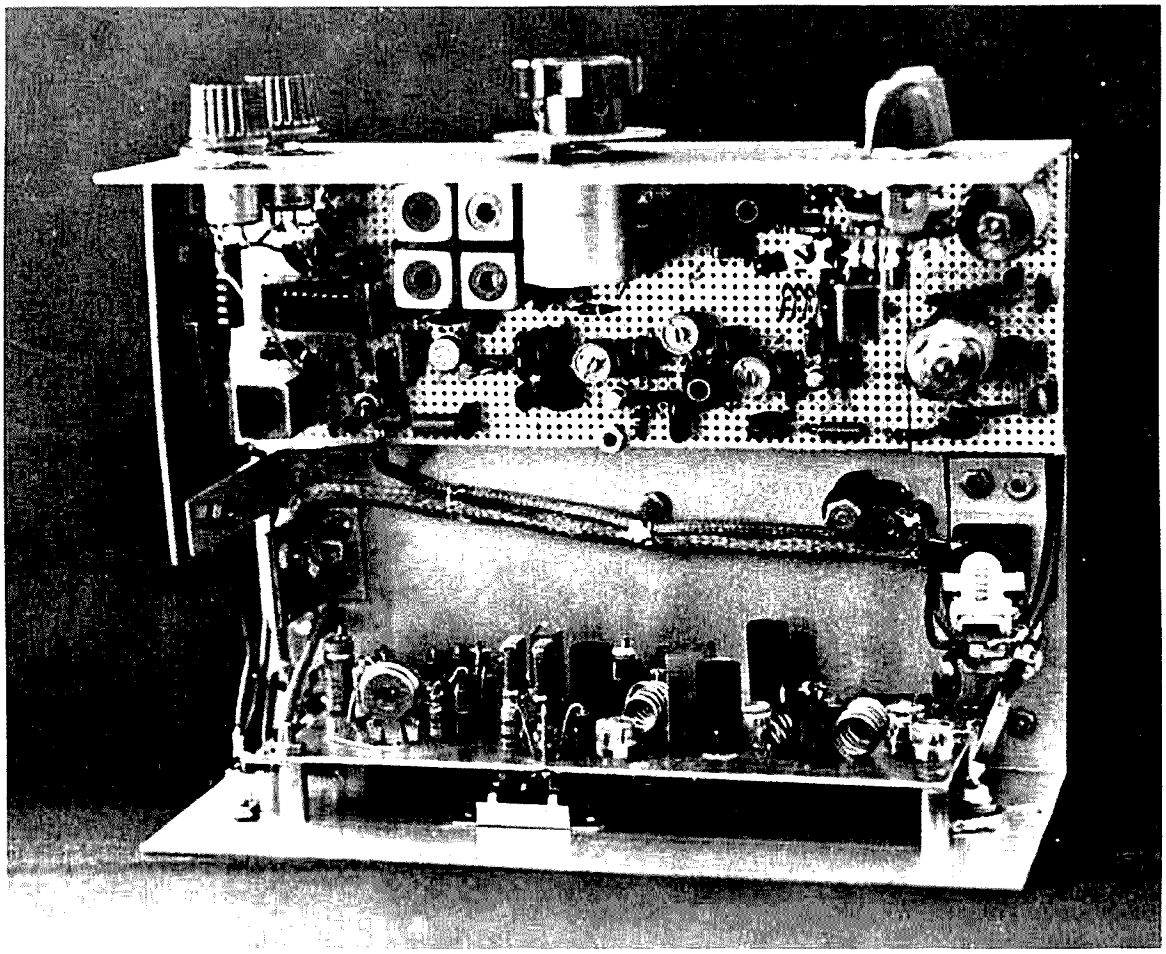


ply is about 1½ watts. This is sufficient to reliably trigger a local repeater from the mobile 10 miles away.

While the transmitter shown in the photographs uses only two crystals operated by a slide switch, it would be equally satisfactory to omit the crystal sockets on the PC board and replace the slide switch with a single external socket for use with any crystal you wish. Tuning can be compromised to use crystals whose two-meter outputs are 1 MHz apart.

The receiver input stage, consisting of transistors Q1 and Q2, is a cascode amplifier using fets. This circuit is particularly satisfactory in that it is not critical as to component values, other than the input and output inductors, it is completely stable and gives a voltage gain of 20 to 25 dB. The inductor shields, not intended to be helical resonators, are thin-wall brass tubes with 6-32 nuts soldered to the tops to permit using 3/16-inch tuning slugs for trimming to frequen-





Construction of the two-meter package. The transmitter section is mounted along the rear wall of the enclosure (bottom). The tunable receiver is laid out on perf board and mounted toward the front of the chassis (top). Transmit/receive relay is to the right.

cy. The tubes themselves are secured by wire pins, passing through the PC board and soldered to the foil side beneath.

The PC board is a standard Vector product, pre-punched in a rectangular grid of 0.10-inch spacing between holes, coated with 2-ounce copper foil on one side. This board is comparatively inexpensive and lends itself to circuit carving with high speed hand tools such as those made by Dreml, Casco, etc. The only drawback is that the board is only 6½-inches long, which required patching on the additional length required. The attractive feature of this type of PC board is that parts may be laid out on the top surface, adjusted to existing holes, interferences resolved and the necessary circuitry drawn in pencil first and then transferred to the foil side for engraving.

Returning to the circuitry, I decided that the second i-f should be somewhere near the top of the broadcast band for reasons which will be discussed later. Since I had also decided that continuous tuning of the entire two-meter band was desirable and practical, it quickly became apparent that 500-kHz segments would be a reasonable target if a medium high first i-f frequency were chosen.

In order that no harmonics of the second conversion oscillator fall into the two-meter band, it develops that the lowest tuning range of the second conversion oscillator must be in the 11.385 to 12.000-MHz range (13th harmonic of 11.385 MHz = 148 MHz, 12th harmonic of 12.000 MHz = 144 MHz). If the second oscillator is set to cover 11.440 to 11.940 MHz, the median frequency is 11.690 MHz. Subtracting the second i-f gives a median frequency of 10.230 MHz for the first i-f. Therefore, the first i-f bandpass range is 9.980 to 10.480 MHz which is comfortably clear of the commercial fm tuners i-f of 10.7 MHz.

This juggling of frequencies does not come by divine inspiration or revelation but happily turns out to be an acceptable solution which does not require extreme measures in the way of second oscillator stabilization.

With the first i-f established at 10.230 MHz and covering plus and minus 250 kHz, you can pick the frequencies for your crystals for the first conversion oscillators. For example, if you want to cover the band from 147.00 to 147.50 MHz, take the median frequency of 147.250, subtract the median i-f frequency of 10.230 MHz to get the first oscillator injection frequency of 137.020 MHz. This means you must order a third-overtone 47.673-MHz crystal and triple that to feed into the first mixer.

The first oscillator circuit is conventional, with its collector tuned to the third overtone of the crystal and driving the base of a second transistor operating as a tripler. Using a second transistor gives a cleaner injection signal and insures adequate drive for the mixer. The dual-gate mosfet mixer is greatly preferred over plain fet mixers with regard to overall performance and the amount of drive required. Tuning the oscillator tripler circuit permits establishing the optimum amount of required injection drive, and the mixer output can be significantly improved by such tuning.

The first mixer feeds the second mixer input through a pair of bandpass circuits. The second mixer is also a mosfet. Its output is tuned to the second i-f frequency of 1460 kHz. The output impedance is stepped down by a capacitive divider to match the low input impedance of the CA3028A IC. Drive to the CA3028A is applied through a series tuned circuit which provides the desired low impedance drive and also discriminates against any first i-f and second oscillator signals appearing in the second mixer drain circuit.

variable oscillator

The second oscillator, covering the tuning range of 11.440 to 11.940 MHz, was one of the major hurdles of the whole design. Capacitance tuning was tried, rather pessimistically, and I was not disappointed. As expected, it was noisy, tuning was much too critical and the capacitor occupied too much space. Next, I tried slug tuning the inductor with various slugs, both powdered iron and ferrite. This eliminated the noise but, with any reasonably compact coil, would not cover the desired tuning range.

Finally, a small Millen coil from the junk box was tried. This is a ceramic coil, 13/32-inch in diameter, ½-inch long with a threaded mounting boss, and tuned with a silver-plated slug, the threaded screw of which projects through the mounting boss (Millen part number 69041).

This coil form turned out to be a prime solution to the tuning problem. The required 500 kHz can be tuned in three turns, which gives adequate spread, and if the slug is properly positioned in the winding, tuning is nearly linear. (This is with the external 1-inch aluminum shield shown in the photographs. Without the shield, the tuning rate is much faster.) As indicated in the schematic, the collector voltage of this oscillator is zener regulated, using the base-emitter junction of a 2N3392 which regulates at 8.6 volts.

The maximum drift of the entire receiver, after two hours operation, is less than 20 kHz, i.e., 10 units on a 100 unit dial. How much of this is attributable to the tunable oscillator I don't know, and quite frankly, don't care. The tuning is smooth, absolutely noiseless and perfectly retraceable. The slug travel is established by the mechanical limit in the coil form at one end, and by positioning and locking a ¼-inch bakelite rod threaded on the slug screw. The tuning dial is on the bakelite rod.

The 1460-kHz output of the second mixer drives a CA3028A in cascode configuration strictly for gain purposes, which in turn feeds a 4-pole filter for high selectivity. This filter consists of four Miller pot-core miniature rf inductors. These inductors are available in a series of ranges, the type 9054 being the most suitable. The aluminum shields on

these inductors must be removed to connect a miniature ceramic capacitor across the winding to provide a shielded tunable circuit. The shields must, of course, be replaced. This requires some study and care in handling. If the Miller 9054 is not available, other inductors can be used with appropriate size capacitors to insure that they can be tuned to the 1460-kHz i-f.

detection and audio output

From the above filter the signal goes to the CA3089 IC. This IC has been both touted and castigated. It is a multipurpose device of great potential, but is extremely intolerant of careless circuit board layout. Briefly, it can provide hard limiting for fm purposes, considerable gain, agc, afc, quadrature fm detection, audio output and – a violent headache. It also has a squelch output and provision for driving a tuning meter. The squelch output is very effective in handling ignition noise when operating mobile. Considerable information, in detail, on the CA3089 is available on request from RCA. It is recommended that anyone contemplating using this IC take advantage of this material.

While the CA3089 was developed for commercial fm reception, it seemed the better part of valor not to attempt using it at the 10.7-MHz level, in view of the reported difficulties in taming it. On the other hand, the internal capacitances in the device which sum up the limiting control are so small that I felt it would be self-defeating to try to use it at a frequency as low as 455 kHz. Hence, the choice of the top of the BC band for the second i-f. At this frequency CA3089 performance, while a little ticklish, is all that could be wished for. Limiting is good, the audio recovery of narrowband fm is excellent and the tuning meter output, either voltage-wise or current-wise, is as described.

The audio amplifier is the LM380 IC made by National Semiconductor. This is a 14-pin DIP that will give nearly 1.4 watts output with a 12-volt power supply. No peripherals are required other than signal input and output and power-supply connections! The LM380 has an optional connection for high frequency bypass and the device will drive an 8-ohm speaker through a capacitor. It does require a 6-square-inch heatsink for its full rating of 2 watts. The photographs show the heatsink (a piece of copper-clad phenolic board) with the IC mounted on it at the end of the receiver PC board. The coupling capacitor is mounted directly on the speaker. The miniature phone jack mounted on the heatsink board permits using either the speaker in the cabinet or a remote speaker. Of course, a remote speaker will require its own coupling capacitor.

With regard to additional items not mounted on the transmitter or receiver boards, there is a miniature closed-circuit phone jack below and behind the heatsink. This is a practical necessity for adjusting and loading the transmitter when first put into operation. A 500-mA meter on a miniature phone plug should be used. Above the meter jack is the power input jack to take a plug from either a car battery or from an ac-operated power supply. Below the power jack is the three-connector microphone jack. The lead from the audio input to the transmitter board is less than 1-inch long.

The push-to-talk microphone lead transfers the input power from receiver to transmitter via the relay seen at the other end of the cabinet. This relay is a two-pole double-throw affair salvaged from some surplus equipment, and is used both as the power transfer relay and the antenna switching relay. It is similar to the type (and size) of those found in Command Sets. Although it is stamped, "28 Vdc," it has a 300-ohm coil which indicates that it would be easy on the power supply. By adjusting the fixed stop on the armature to reduce the air gap somewhat, and by adjusting the relay contacts to make lighter pressure, it was possible to make this relay operate on 10.5 volts. This is not unique with these relays, others have been similarly treated. It is also possible to shim up the contact assembly on some types to reduce the

armature air gap even more. This solved the problem of a suitable relay and the price was right. There is a fixed 100-ohm resistor connected in parallel with the receiver power input to equalize the loads in the transmit and receive positions when using a zener-regulated power supply, but it is not essential.

No-signal power drain is 60 mA at 12 volts in the receive position, kicking up another 100 mA or so with strong signal input. In the transmit position at 12 volts current drain, including relay current, is 350 to 425 mA. This economical power requirement is within the capability of portable battery supply.

converter

I have built a converter for a friend to be used with an automobile BC radio, using the rf amplifier, first mixer and first oscillator to give an i-f output in the 550- to 1550-kHz range. The converter board also contained a broadly tuned i-f transformer from a transistor radio which could be peaked to the approximate i-f, while the car radio is used as a tunable i-f and audio amplifier. With this arrangement two crystals can be used to cover half the two-meter band, the other half of the band appearing as a simultaneous image, not so confusing as might appear at first blush. One set of signals tunes up in frequency while the other tunes down. This unfelicitous arrangement comes out with a measured sensitivity of 2 microvolts, (lacking somewhat in selectivity) while the complete receiver comes off with a measured sensitivity of 0.8 microvolt.

alignment

Receiver alignment can be done with a CW signal except for final adjustment of the quadrature coil. First, disable both oscillators and open the series circuit to the CA3028A input. Feed a low-level 1460-kHz signal to the choke input to the CA3028A. Using a voltmeter on the CA3089 meter lead as indicator, peak the four inductors of the second i-f filter and adjust for equal response roll-off each side of the peak. Reconnect the series input to the CA3028A and shift the 1460-kHz signal input to the vicinity of the second mixer input and tune the second mixer output for maximum indicator response.

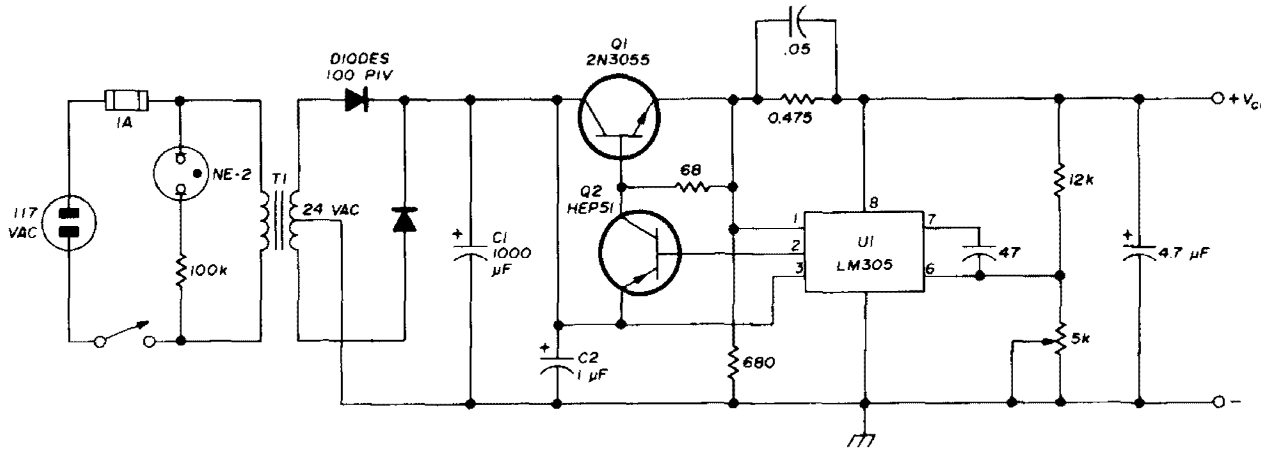


fig. 3. Ac power supply for operating the two-meter package at the base station.

Remove the short from the second oscillator coil and tune the second oscillator to 11.690 MHz. Couple a low-level 10.230-MHz signal to the vicinity of the first mixer input and tune both bandpass circuits for maximum response. It may be necessary to resistively load one circuit for smooth response curve (don't forget to remove the loading when alignment is completed). Restore first oscillator operation and provide an appropriate two-meter signal to the receiver input.

Peak the output circuit of the cascode amplifier and then the input circuit. The former will peak sharply while the latter will have broad peak. Return to the first

oscillator and adjust the tripler circuit to vary injection voltage to the first mixer for best response to a medium-strength signal. Finally, tune in an fm station on two meters and adjust the quadrature coil for best undistorted audio recovery. If you get on an active repeater frequency, you will have a variety of signal qualities to pick from.

ac power supply

A suitable power supply for fixed station use is shown in fig. 3. The LM305 voltage-regulator IC is in a TO-5 can. It provides excellent regulation and current limiting. With the value of R3 indicated it will limit at about 1 ampere. The 2N3055 should have a heatsink and it is desirable that Q2 and the IC also have heatsinks. The tantalum capacitors are specified as such by National Semiconductor as safeguards against parasitics in either the power supply or the load. With the component values shown in fig. 3 the supply will provide 400 mA at 12.3 volts. The output voltage may be reduced considerably, for preliminary test purposes, by means of the 5k pot. All the components for the power supply, except the transformer and 1000-μF filter, can easily be accommodated on a 3-inch-square board.

summary

I don't expect that many amateurs will undertake a project of this magnitude on faith alone, but, hopefully, many readers can use some parts of the design. Some experience with the use of the high-speed carving tools is an absolute necessity.

One of the major problems of compact construction is the source of components. This applies particularly to the physically small items. One prolific source is surplus boards – assuming that the items are quickly identifiable by the seeker. The best source, by far, for new components is Newark Electronics in Chicago. However, you must have their industrial catalog available for ordering parts, and this approach is not inexpensive.

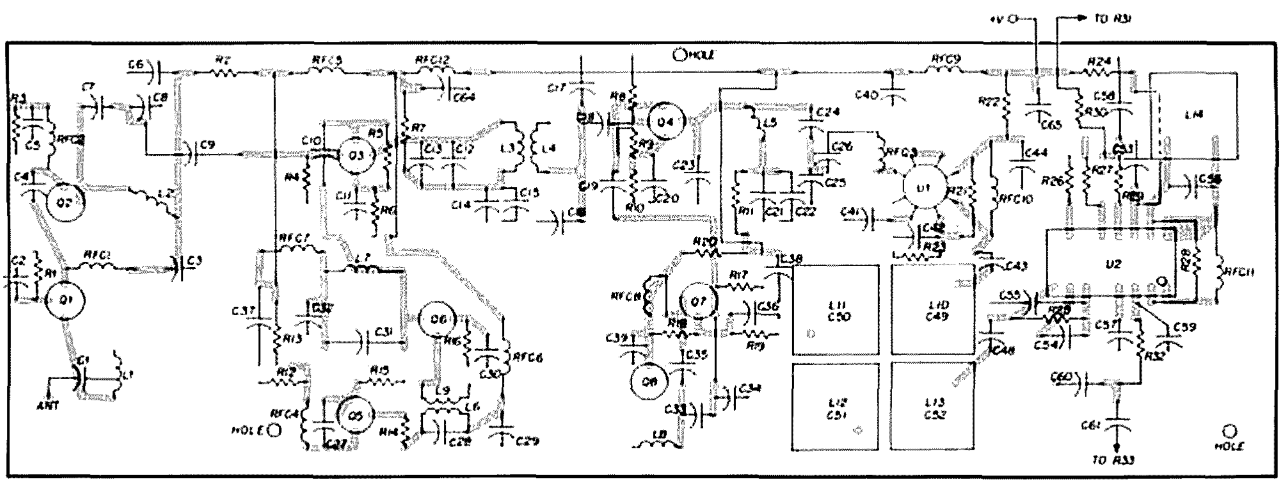


fig. 4. Printed-circuit layout for the tunable two-meter fm receiver. This layout is approximately two-thirds actual size.

There is considerable satisfaction in operating equipment using solid state devices. In addition to compact size, light weight and low power requirements, there is the assurance that, barring mechanical damage, the performance will remain undiminished for the indefinite future. I have been very well satisfied with the performance of this equipment pending the time when some inventive soul produces a 10-watt vhf power transistor in a TO-5 can with integral heatsink. Then back to the drawing board!

reference

1. Doug DeMaw, W1CER, "The Pip-Squeak Gets Smaller," *QST*, September, 1972, page 37.

ham radio

how to solve transistor heatsink problems

A complete discussion of transistor heatsinks and how to choose the one you need for your application

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

How much heatsink is enough? That's a question which can be answered by whether or not the transistor survives when power is applied to the circuit, but this is an expensive way of finding the solution to heatsink design problems; a better way is to calculate heatsink requirements, and select a suitable heatsink using manufacturers' specifications, or make your own heatsink based on empirical data.

A heatsink problem is just one part o' the larger problem of designing a tran sistor circuit to do a particular job; afte the performance requirements of the cir cuit have been established, answers to the following questions must be found. How much power must the transistor dis sipate? Will the selected transistor dis sipate this much power? If so, how much heatsinking is required?

estimating dissipation requirements

Power dissipated by the transistor is that power which is wasted in heating the transistor; it is equal to the power de livered to the transistor minus the power the transistor delivers to its load. A typical example is shown in **fig. 1.** The total power into the transistor is the sum of the signal power into the base plus the dc power delivered to the transistor by the power supply; power wasted in bias resistors does not count. Power delivered by the transistor to its load is the power into the output matching network.

Let's put some hypothetical numbers on the circuit of **fig. 1.** Let's say the amplifier is operated class A, dc collector current is 200 mA, V_{CC} is 12 volts, and the emitter voltage, V_E, is 2 volts. Let's further assume that there is no dc voltage drop in the primary of the output matching network. Therefore, the dc voltage from collector to emitter is equal to 12

minus 2, or 10 volts. With 200 mA of collector current, the dc power into the collector is

$$P_C = V_{CE} I_C = (10 \text{ volts})(0.2 \text{ amp}) = 2 \text{ watts}$$

If the dc beta, or current gain, of the transistor is 50, then the base current is

$$I_B = \frac{I_C}{B} = \frac{200 \text{ mA}}{50} = 4 \text{ mA}$$

Assume the base-emitter voltage of the transistor is 0.6 volt; then the dc power into the base will be

$$P_B = I_B V_{BE} = (4 \text{ mA})(0.6 \text{ volt}) = 0.0024 \text{ watt}$$

This is insignificant compared to collector power and may be neglected.

Assume the transistor delivers 0.9 watt of ac power to the output matching network, and the transistor has a power gain of 10. In this case the signal power into the transistor is

$$\frac{0.9}{10} = 0.09 \text{ watt},$$

and total power into the transistor would be 2.09 watts. The transistor must dissipate the difference between total input power and output power, or

$$2.09 - 0.9 = 1.19 \text{ watt}$$

When the input signal is removed, however, the transistor must dissipate 2 watts because none of the power from the collector power supply goes into the load.

Class-B and class-C amplifiers can be handled in much the same way, the important difference being that the transistor does not dissipate appreciable power when the input signal is removed. Transistor dissipation will still be approximately equal to the difference between the dc power furnished by the collector power supply and the signal power delivered to the collector load.

Fig. 2 shows an example of a class-C rf amplifier. The dc power from the power supply is 2.4 watts, and the power into the T matching network is 1.5 watts. Therefore, the power which must be dissipated in the collector of the transistor is 2.4 minus 1.5 or 0.9 watts. If the power gain of the transistor is 10, then 0.15 watt must be fed into the base of

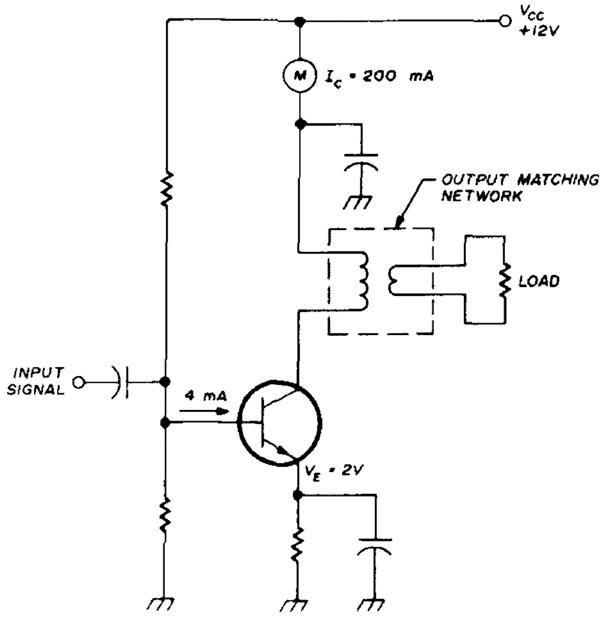


fig. 1. Class-A transistor power stage used to illustrate power dissipation (see text).

the transistor, making a total of 0.9 plus 0.15, or 1.05 watt, which must be dissipated by the transistor.

Transistors used in power supply regulators often require heat sinks. **Fig. 3** shows a simple regulator which delivers 12 watts to the load. Neglecting the transistor base current and zener diode current, the power into the regulator from the rectifiers is

$$18 \text{ volts} \times 1 \text{ amp} = 18 \text{ watts}$$

The transistor must dissipate the difference between input power and output power, or 6 watts.

transistor capabilities

After the power dissipation is estimated, a transistor may be selected which meets the power requirement; this is done by studying the data sheets of transistors

which meet other circuit requirements such as gain and frequency range.

Most transistor data sheets show one or both of the following power ratings. One is total device dissipation at (or below) 25°C *free-air* temperature; this is to simply mount the transistor on a heatsink and allow air at room temperature (25°C) to circulate around it. This method will not hold the case temperature at 25°C, so a reduced power rating must be accepted for the device.

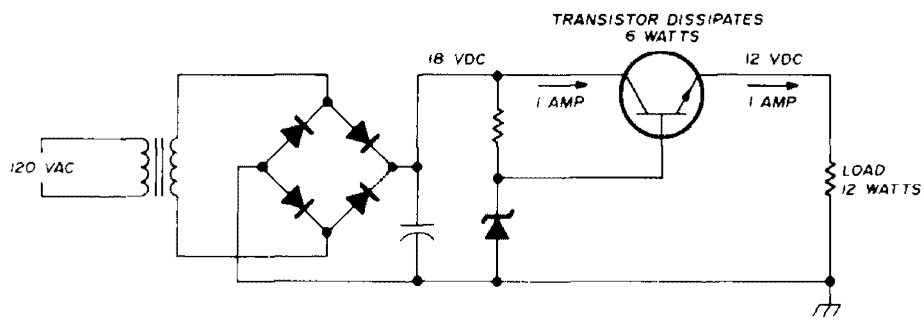


fig. 3. The series regulator transistor in this simple power supply dissipates 6 watts.

the maximum power the transistor may dissipate without any heatsink if the temperature of the air around the transistor is 25°C or less.

The other rating is total device dissipation at (or below) 25°C *case* temperature; this is the maximum power the transistor may dissipate if the transistor case temperature is held to 25°C or less. One way to hold the case temperature at 25°C is to mount the transistor on a heatsink which has integral cooling coils through which ice water is pumped. This is rather expensive, however, and the usual procedure is

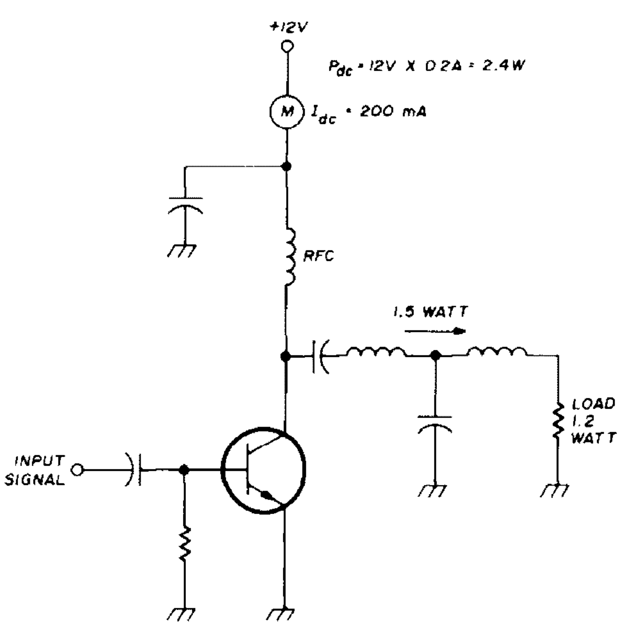


fig. 2. The transistor dissipates 1.05-watt in this class-C rf amplifier stage.

Included with the free-air and case-temperature ratings are derating factors which say that the device must be derated linearly to some temperature (free-air or case) at the rate of so many watts-per-°C. Derating factors may be shown as footnotes on the data sheet.

Some transistor data sheets include dissipation derating curves which are merely graphs showing what has already been specified; if the data sheet does not include it, one may easily be drawn. **Fig. 4** shows derating curves for the 2N3724A transistor. This device is rated for 1-watt dissipation at 25° C free-air temperature (no heatsink) with a derating factor of 5.71 mW-per-°C to 200°C. Notice that if the free-air temperature is 200°C, no power may be dissipated by the transistor. If the free-air temperature is 50°C, then the amount of derating is 25°C times 5.71 mW-per-°C, or 143 mW; this derating is subtracted from the 1-watt rating to find how much power may be dissipated at 50°C free-air.

$$1000 \text{ mW} - 143 \text{ mW} = 857 \text{ mW}$$

The case temperature curve applies if a heatsink is used. If the transistor dissipates, say 3 watts, then, according to **fig. 4**, the heatsink must be large enough to hold the case temperature to 95°C or less. Operation must always be on the curve or below it. It is good design practice to

allow some safety factor by operating somewhat below the curve, i.e., use a slightly larger heatsink than is called for. As mentioned above, practical heatsinks will not hold the case temperature to 25°C, so you should not expect to operate this device at 5-watts dissipation.

Some transistors have power ratings specified at 50°C or 100°C case temperatures, with appropriate derating factors for case temperatures above those values. Fig. 5 shows the derating curve for the 2N5387. This transistor is rated for 100 watts at (or below) 100°C case temperature and has a derating factor of 1 watt-per-°C to 200°C case temperature. It is possible for transistors with such ratings to dissipate their full power rating using an air-cooled heatsink, provided the heatsink is good enough.

thermal resistance

Thermal resistance is expressed in the units, °C-per-watt. This is the temperature difference that will occur between two points for each watt of power that is dissipated at one of the points, the higher temperature being at the point where power is dissipated. The reciprocals of the derating factors discussed above are thermal resistances.

$$\text{Thermal Resistance } \frac{°C}{\text{watt}} = \frac{1}{\text{Derating Factor } \frac{\text{watts}}{°C}}$$

Theta (θ) is the mathematical symbol used for thermal resistance, and subscripts are used to denote which two points the thermal resistance is between:

$\theta_{J\text{-}A}$ junction-to-ambient
$\theta_{J\text{-}C}$ junction-to-case
$\theta_{C\text{-}HS}$ case-to-heatsink
$\theta_{HS\text{-}A}$ heatsink-to-ambient

Junction is the term used for the point or points inside the transistor where the power is actually dissipated. Case means the point or points on the transistor package where the heatsink makes contact. Ambient is the medium into which heat is ultimately conducted or radiated, and it usually is free-air at 25°C.

If the transistor or heatsink is mounted inside an equipment cabinet where the temperature is higher than room temperature due to heat-generating components, such as transformers, tubes, power resistors or the power transistor itself, then ambient means the temperature inside that cabinet. **Fig. 6** is a scale showing the relationship between °C and °F for those accustomed to expressing temperature in °F.

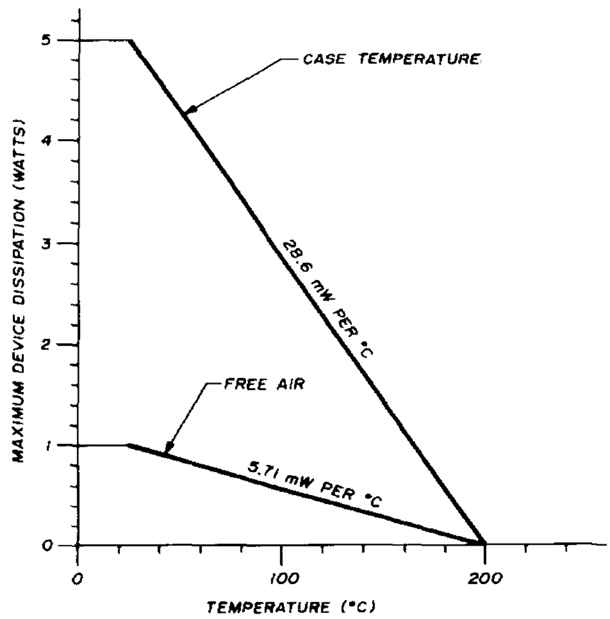


fig. 4. Dissipation derating curves for the 2N3724A transistor.

The reciprocal of the free-air derating factor is $\theta_{J\text{-}A}$

$$\theta_{J\text{-}A} = \frac{1}{\text{Free-air Derating Factor}}$$

For the 2N3724A (see **Fig. 4**),

$$\theta_{J\text{-}A} = \frac{1}{5.71 \frac{\text{mW}}{°C}} = 0.175 \frac{°C}{\text{mW}}$$

$$= 175 \frac{°C}{\text{watt}}$$

This thermal resistance value tells you how many °C the junction temperature will rise above ambient temperature for a

given transistor power dissipation. If the ambient temperature, T_A, is 25°C, and 1 watt of power, P, is dissipated at the junction, then the junction temperature, T_J, will be

$$T_J = T_A + P\theta_{J\text{-}A} \qquad (1)$$

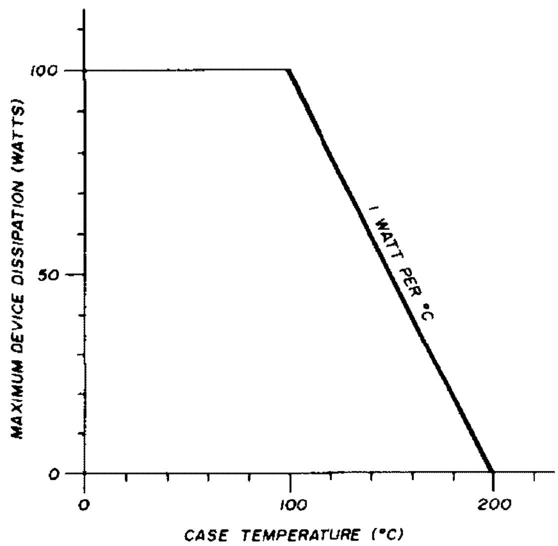


fig. 5. Dissipation derating curve for the 2N3724A transistor.

$$T_J = 25 + (1\ \text{watt})(175\ \frac{°C}{\text{watt}}) = 200°C$$

It is obvious from this and **fig.** 4 that the maximum allowable junction temperature for the 2N3724A is 200°C. *The entire purpose of heat sinking is to prevent the junction temperature from exceeding the maximum allowable value specified by the manufacturer.*

Eq. 1 is the basic thermal equation used to determine if a certain power

When the transistor is fastened to a heatsink, $\theta_{J\text{-}A}$ breaks down into three quantities;

$$\theta_{J\text{-}A} = \theta_{J\text{-}C} + \theta_{C\text{-}HS} + \theta_{HS\text{-}A} \qquad (2)$$

Combining **eqs. 1** and **2** gives

$$T_J = T_A + P(\theta_{J\text{-}C} + \theta_{C\text{-}HS} + \theta_{HS\text{-}A}) \qquad (3)$$

$\theta_{J\text{-}C}$ is the reciprocal of the case temperature derating factor shown on the transistor data sheet;

$$\theta_{J\text{-}C} = \frac{1}{\text{Case Temperature Derating Factor}}$$

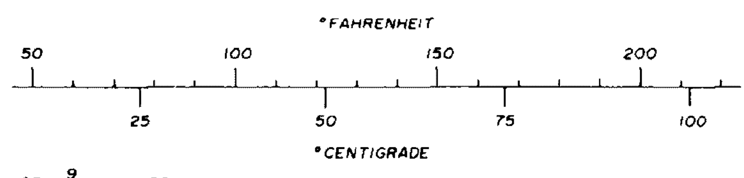
For the 2N3724A (see **fig. 4**),

$$\theta_{J\text{-}C} = \frac{1}{28.6\ \frac{mW}{°C}} = 0.035\ \frac{°C}{mW} = 35\ \frac{°C}{watt}$$

This value of thermal resistance tells you that the junction temperature will be 35°C higher than the case temperature for each watt of power dissipated in the junction. If 5 watts of power are dissipated in the junction, then the junction temperature will be 175°C higher than the case temperature. Therefore, the case temperature *must not* exceed 25°C if the junction temperature is not to exceed 200°C, its maximum allowable value.

Case-to-heatsink thermal resistance, $\theta_{C\text{-}HS}$, depends on several varying factors. How much torque is used in tightening the nuts or screws which hold the transistor to the heatsink? How smooth

fig. 6. Relationship between the Centigrade and Fahrenheit temperature scales.



dissipation will cause the transistor's maximum allowable junction temperature to be exceeded. T_A should be the highest actual ambient temperature encountered. Don't use T_A = 25°C if the transistor is to be operated in the trunk of a car on hot summer days; 50 to 75°C would be more realistic.

are the mating surfaces of the transistor and heatsink? Is the heatsink anodized? Is an insulating mica washer used between the transistor and heatsink? Is a silicone grease, or other thermal compound, applied to the mating surfaces? How and to what extent do these factors affect $\theta_{C\text{-}HS}$? All these factors have an effect.

A mica washer will increase the thermal resistance about 0.3 °C-per-watt, and thermal compounds may decrease the thermal resistance about 0.1 to 0.2 °C-per-watt. Anodized surfaces are about 0.25 °C-per-watt higher than unfinished surfaces. A fair rule-of-thumb is to allow

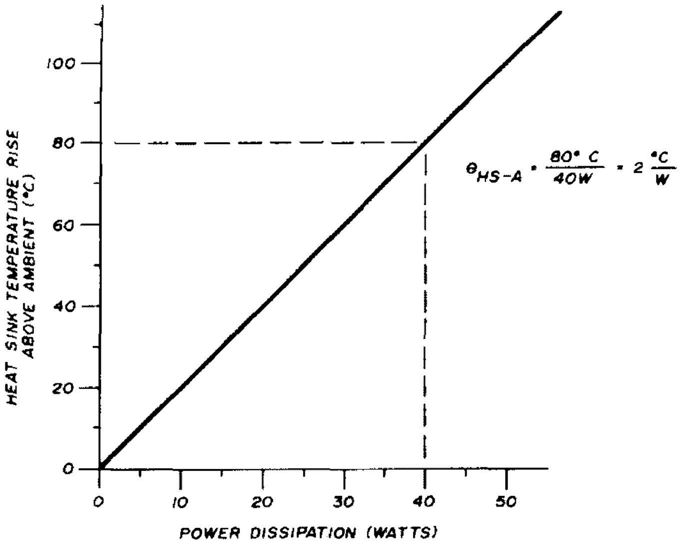


fig. 7. Typical heatsink performance curve.

about 0.2 °C-per-watt for $\theta_{C\text{-}HS}$ when the mating surfaces are bare metal; 0.5 °C-per-watt should be used if the heat sink is anodized or a mica washer is used.

$\theta_{C\text{-}HS}$ is usually not very significant at moderate power levels, but if the power dissipation is 100 watts, the temperature difference between the transistor case and the heatsink could be on the order of 50°C. Mating surfaces should be smooth and clean, and the transistor should be mounted tightly to the heatsink.

The last term in **eq. 3**, $\theta_{HS\text{-}A}$, is the heatsink-to-ambient thermal resistance. Heatsink manufacturers may specify the thermal resistance of their heatsinks in °C-per-watt, or they may provide a performance graph such as shown in **fig. 7**. Since the curve is usually a straight line, the slope of which is thermal resistance, $\theta_{HS\text{-}A}$ may be derived from the curve as shown. The manufacturer may label the vertical axis in **fig. 7** case temperature rise above ambient in °C; in this case $\theta_{C\text{-}HS}$ is included in the heatsink rating, so the slope of the line is equal to $\theta_{C\text{-}HS}$ + $\theta_{HS\text{-}A}$.

finding the right heatsink

To choose a suitable heatsink it is necessary to determine the value of $\theta_{HS\text{-}A}$ you need, then select a heatsink having that value, or less, of thermal resistance. Some examples will illustrate.

Suppose a circuit includes a TIP29 power transistor; the maximum power, P, which the transistor must dissipate is 15 watts. To allow for operation in non-air-conditioned places on hot summer days, ambient temperature, T_A, is assigned a value of 50°C. The TIP29 data sheet specifies that the maximum continuous device dissipation at (or below) 25°C case temperature is 30 watts, and this rating is to be derated to 150°C case temperature at the rate of 0.24 watt-per-°C. Thus, the maximum allowable junction temperature, T_J, is 150°C, and the junction-to-case thermal resistance is

$$\theta_{J\text{-}C} = \frac{1}{0.24 \frac{\text{watt}}{°\text{C}}} = 4.17 \frac{°\text{C}}{\text{watt}}$$

It is desired to use an insulating mica washer when mounting the TIP29 to its heatsink, so $\theta_{C\text{-}HS}$ is assumed to be 0.5°C-per-watt. All of this information is substituted into **eq. 3** as follows

$$T_J = T_A + P(\theta_{J\text{-}C} + \theta_{C\text{-}HS} + \theta_{HS\text{-}A})$$

$$150 = 50 + 15(4.17 + 0.5 + \theta_{HS\text{-}A})$$

Solving for $\theta_{HS\text{-}A}$,

$$\theta_{HS\text{-}A} = 2 \frac{°\text{C}}{\text{watt}}$$

Thus, an acceptable heatsink would be one which has a thermal resistance of 2 °C-per-watt or less. Armed with this information, you can quickly select a suitable heatsink from the manufacturers' catalogs. The Thermalloy Company's 6123 heatsink is rated at 1.3 °C-per-watt and would be quite adequate.

As a second example, assume a 2N5387 (see **fig. 5**) must dissipate 100 watts, and the ambient temperature is 25°C. From **fig. 5**, or from the derating information, the maximum allowable junction temperature is 200°C, and the

junction-to-case thermal resistance is

$$\theta_{J\text{-}C} = \frac{1}{1\ \frac{\text{watt}}{^{\circ}\text{C}}} = 1\ \frac{^{\circ}\text{C}}{\text{watt}}$$

The transistor is to be mounted directly to the bare metal of the heatsink, so $\theta_{C\text{-}HS}$ is taken to be 0.2 °C-per-watt. Plugging these values into **eq. 3** gives

$$T_J = T_A + P(\theta_{J\text{-}C} + \theta_{C\text{-}HS} + \theta_{HS\text{-}A})$$

$$200 = 25 + 100(1 + 0.2 + \theta_{HS\text{-}A})$$

Solving for $\theta_{HS\text{-}A}$ yields

$$\theta_{HS\text{-}A} = 0.55\ \frac{^{\circ}\text{C}}{\text{watt}}$$

The Thermalloy 6560B heatsink should handle the requirement. It is a black anodized heatsink, but the catalog information indicates that for 100-watts dissipation, the transistor case temperature will be 45°C above ambient. This means

$$\theta_{C\text{-}HS} + \theta_{HS\text{-}A} = \frac{45^{\circ}\text{C}}{100\ \text{watts}} = 0.45\ \frac{^{\circ}\text{C}}{\text{watt}}$$

Putting this into **eq. 3**,

$$T_J = T_A + P(\theta_{J\text{-}C} + \theta_{C\text{-}HS} + \theta_{HS\text{-}A})$$

$$T_J = 25 + 100(1.0 + 0.45) = 170^{\circ}\text{C}$$

which is 30°C less than the maximum allowable junction temperature; it would be unnecessary to remove the anodize where the transistor mounts to the heatsink.

homemade heatsinks

Heatsinks may be improvised by using sheet metal. Weight, volume and shape play some part in heatsink effectiveness, but exposed surface area is the prime factor on which thermal resistance depends. **Fig. 8** is a graph showing approximate $\theta_{HS\text{-}A}$ vs area of one side for 1/8-inch thick square aluminum and copper sheet metal. This data applies to square plates mounted so the plane of the plate is vertical, with the transistor fastened to the center of the plate.

The thermal conductivity of copper is nearly twice that of aluminum which explains why copper gives better results. Brass has a thermal conductivity about one-half that of aluminum, and should be avoided; steel is poor also. Aluminum is the best compromise between performance and cost, and it is widely used.

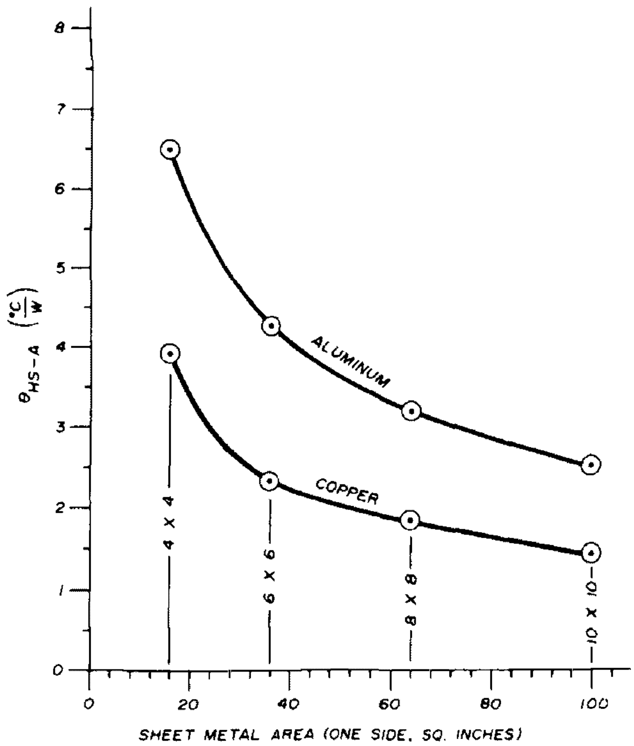


fig. 8. Thermal resistance vs area of 1/8" thick sheet metal.

Fig. 8 may be used to estimate the size of sheet metal needed after the required value of $\theta_{HS\text{-}A}$ has been determined. An aluminum chassis may be used as a heatsink by mounting the transistor to it, but the horizontal portion of the chassis does not get rid of heat as well as the vertical portions. Air rises when it is heated, and all heatsinks should be mounted so most of the surface area is vertical. This permits the most efficient flow of air past the heatsink due to convection currents.

forced air cooling

Blowing air across the surface of a heatsink by means of a fan or blower can dramatically improve the heatsink's performance. For example, air blown at a velocity of 500 feet-per-minute will reduce $\theta_{HS\text{-}A}$ to around one-third to one-

half its still-air value; this corresponds to a light breeze of about 5.7 miles-per-hour.

Fans and blowers are rated in cubic feet-per-minute (cfm). To determine the approximate velocity of air out of a blower, the cfm rating is divided by the cross-sectional area of the blower's output hole. For example, suppose a small blower is rated at 20 cfm, and its output hole is 2 x 2 inches. The cross-sectional area is 4 square inches, or 0.0278 square feet, so the output air velocity is

$$\frac{20\ \text{cfm}}{0.0278\ \text{sq ft}} = 720\ \frac{\text{feet}}{\text{minute}}$$

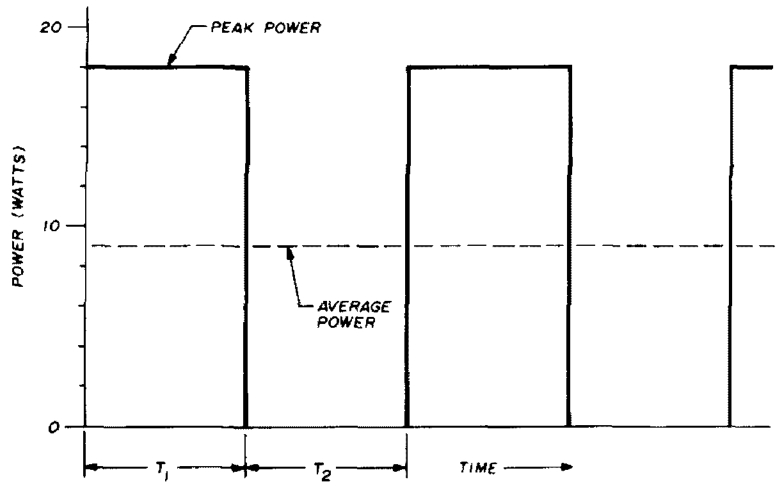


fig. 9. Power dissipated in a junction by a square wave.

If the blower cannot be mounted so that its output flows directly onto the heatsink, ducting may be used to channel the air stream to the heatsink. More detailed information on forced-air cooling may be found in heatsink catalogs.

thermal time lag

A finite amount of time is required for the junction temperature to rise to its steady-state value after power is applied. This fact allows the transistor to operate at higher ac peak power ratings than is possible at dc. **Fig. 9** shows a graph of the power dissipated in a transistor junction by a square wave; peak power is 18 watts, and average power is 9 watts.

Assume that the transistor and its heatsink are such that only 10 watts of power may be safely dissipated. If the frequency of the square wave in **fig. 9** is 500 kHz, T_1 and T_2 will each be 1 microsecond. The thermal time constant of most power transistors is large compared to 1 μs, so the junction temperature will be determined by the average power dissipation of 9 watts, and the junction will operate within its rating.

Now suppose the frequency of the square wave is decreased to a very low value, say one cycle-per-hour; T_1 and T_2 would each be 30 minutes, and 18 watts would be applied to the junction for 30 minutes during the first cycle, quickly destroying the device. This is an extreme example, but it points out the necessity of taking frequency into consideration when determining heatsink requirements.

Many power transistors have thermal time constants such that the use of average power in **eq. 3** would lead to an inadequate heatsink at the lower audio frequencies. Therefore, it is not a bad idea to use peak power in eq. **3** for audio frequency applications.

Some transistor data sheets show a family of curves to be used in adjusting the value of $\theta_{J\text{-}C}$ according to pulse width and duty cycle, and these should be studied and used when available.

conclusion

It is hoped that this article will introduce the reader to the basic concepts involved in solving transistor heatsink problems; these principles may also be applied to other semiconductor devices such as thyristors and power zeners. More insight into heatsink technology may be derived by studying power transistor data sheets and heatsink catalogs. **ham radio**

simple lowpass filter for audio

This simple lowpass audio filter provides high performance and a minimum of design effort — a design graph is provided

Frank Regier, OD5CG, American University of Beirut, Lebanon

Lowpass audio filters have many applications in amateur radio, such as restricting transmitter bandwidth and establishing the bandwidth of direct-conversion receivers. Simple tee- and pi-section filters are often used in these applications but do not provide sharp cutoff. The circuit presented here is substantially better than a tee or pi but is nevertheless inexpensive and simple to build.

So-called modern filters are the best that can be made for a given number of components, but these components are likely to have awkward, nonstandard values. The filter to be described here performs very well and is much easier to make than a comparable modern filter. It consists of three unmodified telephone toroids and four identical capacitors — nothing more.

The filter is composed of a constant-k pi-section with an m-derived half-section at each end. For best matching to a resistive load, such half-sections are usually made with m = 0.6. If, however, you let m = 0.5, for only a slight degradation in performance you achieve two important simplifications. First, all capacitors in the circuit assume the same value, and second, each end inductor assumes exactly one-quarter of the value of the center inductor. This latter property makes it possible to use an 88- or 44-mH telephone toroid for the center inductor and half of a similar toroid for each of the end inductors. The resultant filter is shown in **fig. 1.** It is important that it be terminated in its proper load resistance, R.

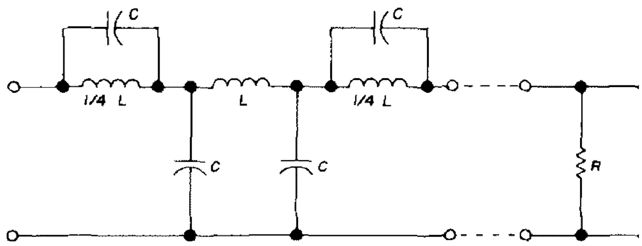


fig. 1. Simple lowpass audio filter uses three inductors and four capacitors. For L = 88 mH, the values of R and C may be obtained from fig. 2.

design

The design of a particular filter begins with the choice of a cutoff frequency. Then, with the value of L known, the values of R and C are obtained from the equations

$$R = \pi L f_c \text{ ohms} \tag{1}$$

$$C = \frac{0.75}{\pi^2 L f_c^2} \text{ farads} \tag{2}$$

where L is in henrys and f_c is the cutoff frequency in Hz.

If 88-mH toroids are used, the values of R and C may be obtained graphically from fig. 2 for a considerable range of cutoff frequencies. For 44-mH toroids, the values of R should be half, and the values of C double, those shown in **fig. 2.**

To test the design, a filter was built using L = 88 mH and C = 0.1 μF. These values lead to a cutoff frequency of 2940 Hz and require a load resistance of 812 ohms. Each of the two 22-mH end inductors was formed by paralleling the two windings of an 88 mH telephone toroid. These toroids have very low core losses at audio frequencies, so their Q is determined almost entirely by winding resistance. The Q of the end inductors can therefore be almost doubled, without changing the inductance, if the two windings are paralleled. For correct polarity, the two braid-covered ends should be joined,and the other two ends should also be joined. The four capacitors were matched to within 1%, and a load resistance accurate to within 1% was made by paralleling higher-value resistances.

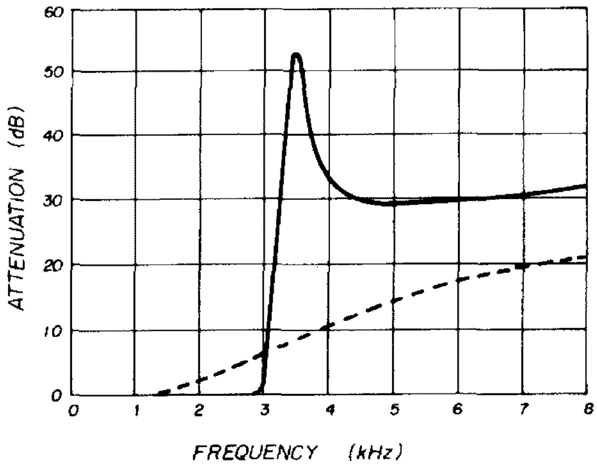


fig. 3. Measured attenuation characteristics of the lowpass filter shown in fig. 1 with a 2940-Hz cutoff frequency. The dashed line shows the measured attenuation characteristics of a comparison pi-section filter with the same nominal cutoff frequency and load resistance.

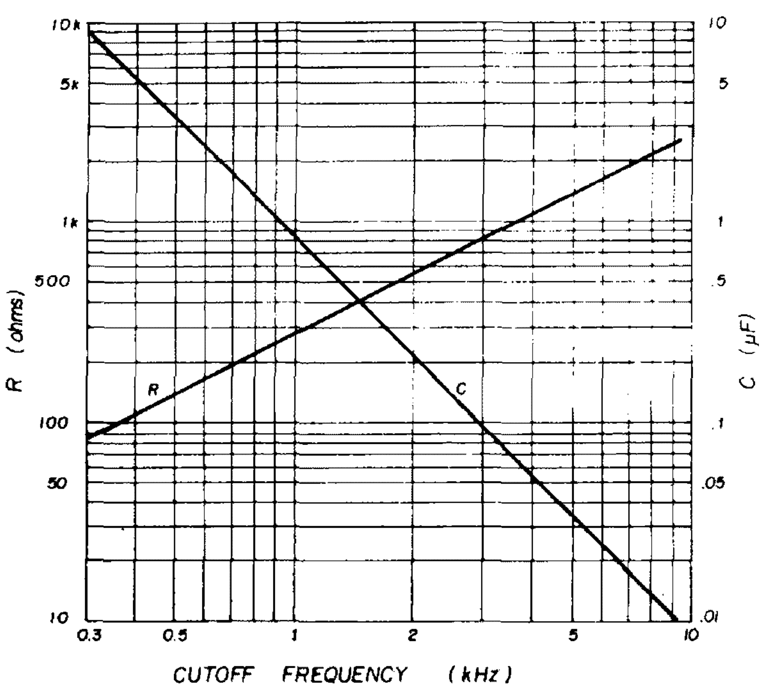


fig. 2. Values of R and C as functions of the cutoff frequency for the lowpass filter circuit shown in fig. 1 with L = 88 mH. For L = 44 mH, C should be doubled and R halved.

operating characteristics

The attenuation characteristics of the resultant filter were measured and are shown in **fig. 3.** Also shown (dashed curve) are the measured attenuation characteristics of a comparison pi-section filter having the same cutoff frequency and load resistance. The pi-section filter was made from an 88 mH toroid and two 0.067 μF capacitors. The superiority of the filter circuit of **fig. 1** is obvious.

ham radio

medium-power toroidal antenna tuner

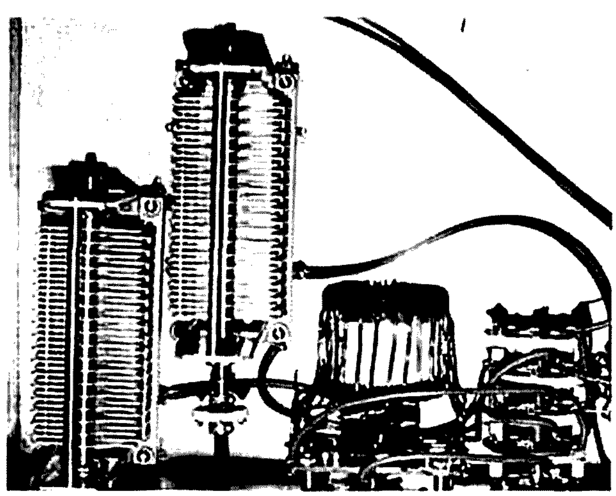
Design and construction of a compact antenna tuner that will handle up to 500 watts

Gregory Widin, WB2ZSH, Box 248, Gambier, Ohio 43022

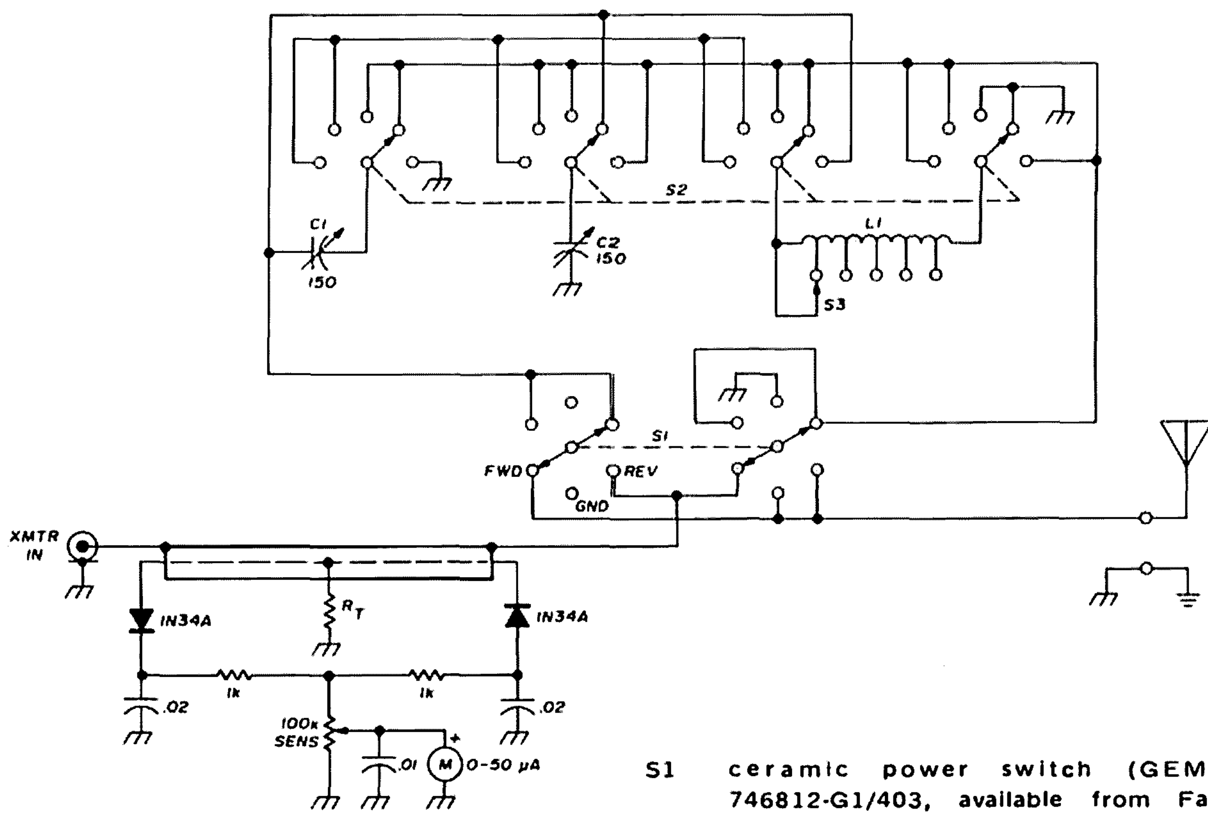
Though numerous designs have been presented for antenna tuners, most are anything but simple and convenient to use. Moreover, the tuners described for limited space applications are themselves often far from compact.

The antenna tuner described here overcomes these problems by attacking the primary culprit — the inductor. Designs using plug-in, rotary or clip-tapped inductors are superseded with the use of a switch-tapped toroidal coil, thereby substantially reducing space requirements and the inconvenience of bulky tap connections. The circuit is based on the recommendations of W2EEY,[1] and provides matching to random length wires. An indicator is included for "hands-free" tuning.

The coupler provides nine different circuits using two capacitors and one coil (see fig. 2). Configurations A through E are provided by switch S2, and the forward/reverse function is accomplished by switch S1. The different circuit arrangements provide for a variety of impedance-matching situations.



Closeup of the matching network components in the toroid antenna tuner. Variable capacitors C1 and C2 are to the left, toroid inductance L1 is to the right.



C1,C2	150-pF variable (Millen 12515)	S1	ceramic power switch (GEML 746812-G1/403, available from Fair Radio)
L1	24 turns no. 12 Formvar wound on two Amidon T-200-2 cores, tapped at 4, 6, 8, 10 and 12 turns (see text)	S2	4-pole, 6-position, non-shorting rotary switch (CRL 2553)
R_T	same resistance as load, 50 ohms for most amateur systems (see reference 2)	S3	12-position tap switch (JBL Inst. 240 type BPBN-1-RA-83-3Z9825-40.2, available from Fair Radio)

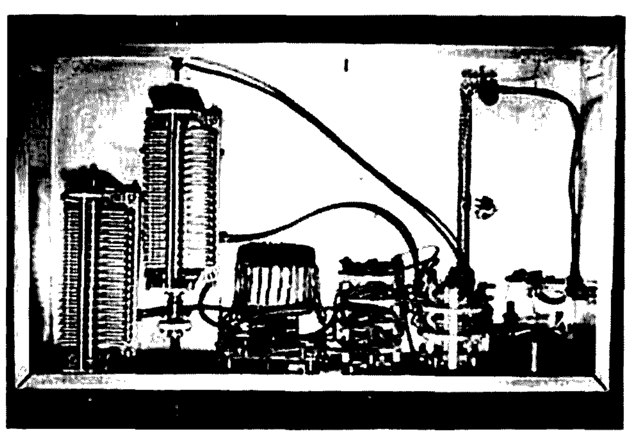
fig. 1. Schematic diagram of the toroid antenna tuner. This tuner will handle up to 500 watts CW without arcing, and is designed primarily for matching long-wire antennas from 80 through 10 meters.

construction

Building the toroidal antenna tuner is relatively straightforward. The toroid is the most unusual part of the circuit and consists of two Amidon T-200-2 toroidal cores epoxied together. The entire surface of each of the toroids is covered with epoxy to prevent flashover from the coil to the cores. Spacers of 1/4-inch polystyrene are then cut out as shown in fig. 3 and glued to the ends of the dual toroid.

When the epoxy has cured, the wire may be wound on the toroids — 24 turns of number-12 Formvar-insulated wire are required. Care should be taken not to flex the wire more than necessary, as this will work harden the wire. Also, the neater the job, the less likely you will have arcing problems in the finished tuner. Leave enough wire at each end to secure the coil to the tap switch.

The tap leads from the coil are connected before the coil is wired to the



Construction of the toroid antenna tuner. All components are mounted in a small aluminum chassis.

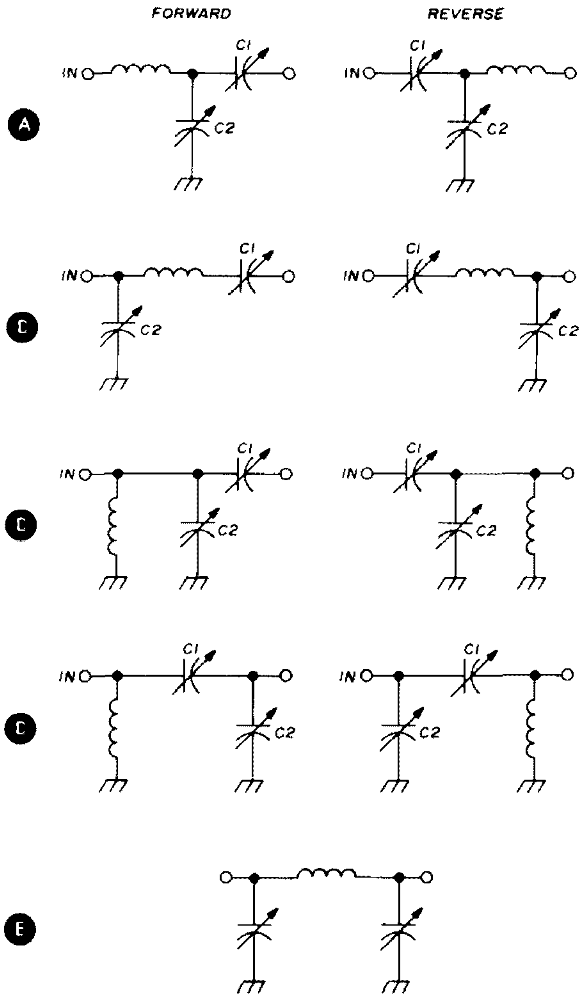


fig. 2. Different matching network arrangements possible with the antenna tuner shown in fig. 1.

switch. Beginning after the first 4 turns, taps are connected every 2 turns, for a total of 12 leads, including those at each end of the coil. To connect the taps, scrape away the insulation on the proper turn on the outside of the coil between the spacers. Another piece of number-12 wire with a clean end is then wrapped to this point with several turns of small

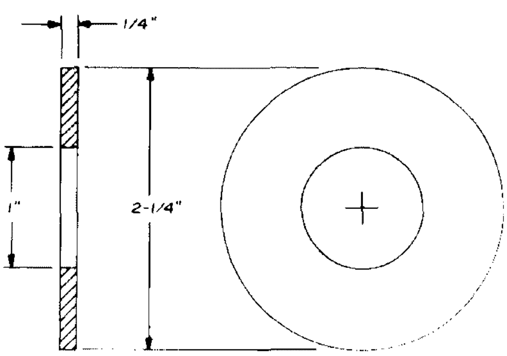


fig. 3. Polystyrene end spacers for the toroidal core (see text).

gauge wire. Then the connection is soldered.

When the tap leads are all connected, the coil may be wired to the switch. The first contact is left blank, and the second contact connects to the first tap after the initial 4 turns on the coil. The wires are connected around, in turn, and support the coil. The wiper should be connected to the end of the coil with the 4-turn tap.

The capacitors were obtained, in new condition, from a surplus A-27 Phantom Antenna unit.* These units also supplied the ground and antenna binding posts, which are more rugged than most. Note that one capacitor must be fully insulated from ground.

Since the forward/reverse switch provides 3 positions, the center position is used to ground the antenna when the equipment is not in use. A dummy load might be connected to the input side in this position to provide a tune option.

The swr indicator shown in **fig. 1** is a modification of an earlier design.[2] I used a pickup braid 8-inches long to give significant indication in the transmitter tune-up position. The sensitivity control used was a subminiature type, but a front-panel adjustment would be more satisfactory. The indicator portion of the tuner should be shielded to prevent possible rfi effects.

The tuner could easily be built into a small enclosure. Front panel space is the main limitation on compactness. The finished unit is capable of handling 500 watts CW without arcing. Using a long-wire antenna of sufficient length, the tuner will easily match transmitter outputs from 80 through 10 meters.

references

1. John J. Shultz, "Random-Length Antenna Couplers," *ham radio*, January, 1970, page 32.
2. Gregory P. Widin, "SWR Bridge," *ham radio*, October, 1971, page 55.
3. E. L. Klein, W4BRS, "The Whole of the Doughnut," *73*, June, 1967, page 6.

ham radio

*A-27 Phantom Antenna units, used, are priced at $2.95 plus shipping (3 pounds) from Fair Radio Sales Co., Post Box 1105, Lima, Ohio 45802.

four-band high-frequency windom antenna

The rebirth of the Windom antenna — a high performance multiband antenna popular in the 1930s

Hal Morris, W4VUO, 354 Krams Avenue, Philadelphia, Pennsylvania 19128

Do you have antenna space limitations? Can't swing a rotary beam? Need a good field-day antenna? Then the old standby, the Windom antenna, may be your answer. It offers four-band operation with a single feedline, and in most cases does not require an antenna tuner.

It's odd how ideas crop up in ham radio and then fade into oblivion. The Windom is a good, simple, multiband antenna system that is unheard of among today's hams. So, let's revive it and simplify the feed system. (This will be old hat to you if you remember when you weren't one of the boys on 75 meters unless you had an RME-45 receiver and a Windom antenna.)

theory of operation

If the impedances present along the length of a half-wave dipole in free space are plotted, the values vary from about 3600 ohms at the ends to 72 ohms at the center. **Fig. 1** is a plot of antenna impedance versus length along a dipole. The center impedance of 72 ohms, coupled with the ease of using coaxial cable, has given rise to the extensive use of low-impedance feedlines and single-band dipoles. Today, open-wire feeders and other than 50- or 72-ohm coaxial feedlines are rare.

However, one way of feeding a dipole with open wire-line is to tap the antenna

equidistant from the center to match the feedline impedance. Fig. 2 illustrates a method of matching 600-ohm line to a dipole. Note that the dipole does not have to be split into two parts with an insulator. This is called the delta match and is used extensively by vhf enthusiasts for matching stacked arrays.

preplanning

Lets calculate the length required for a four-band antenna. Since the highest frequency band, ten meters, will be the most sensitive to antenna length, overall antenna length must be some multiple of a half-wavelength at ten meters. From the handbook formula for long-wire antennas

$$\text{length in feet} = \frac{492\ (N - 0.05)}{\text{frequency (MHz)}}$$

where N is number of half waves.

For an antenna nine half-wavelengths long at 28.9 MHz, the length is slightly more than 152 feet. This is a bit long for 80-meter operation. Plugging in eight half waves and turning the crank gives

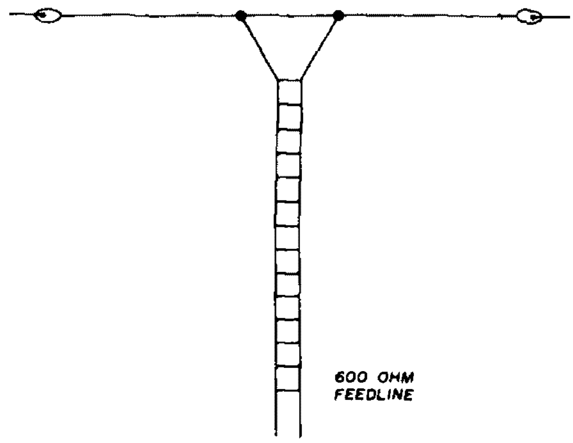


fig. 2. Classic single band antenna uses an open-wire feedline and a delta match. No center insulator is required.

135.342 feet. This looks good. Round the length off to 136 feet.

Now, using the formula for a half-wave dipole, and working backwards to find resonant frequency

$$f_{MHz} = \frac{492}{\text{length}} = \frac{492}{136} = 3.617\ \text{MHz}$$

This looks good. The 80- and 75-meter bandedge mismatch will be a small percentage of antenna length.

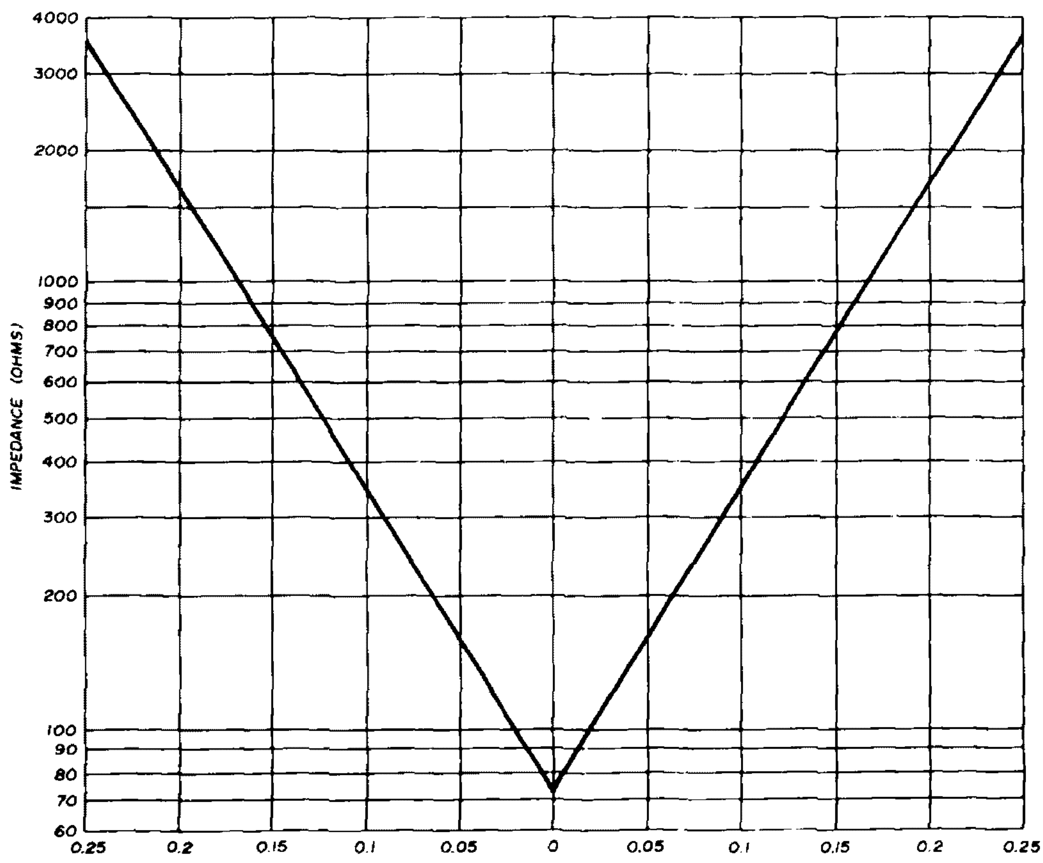


fig. 1. Plot of input impedance along a half-wave antenna in free space.

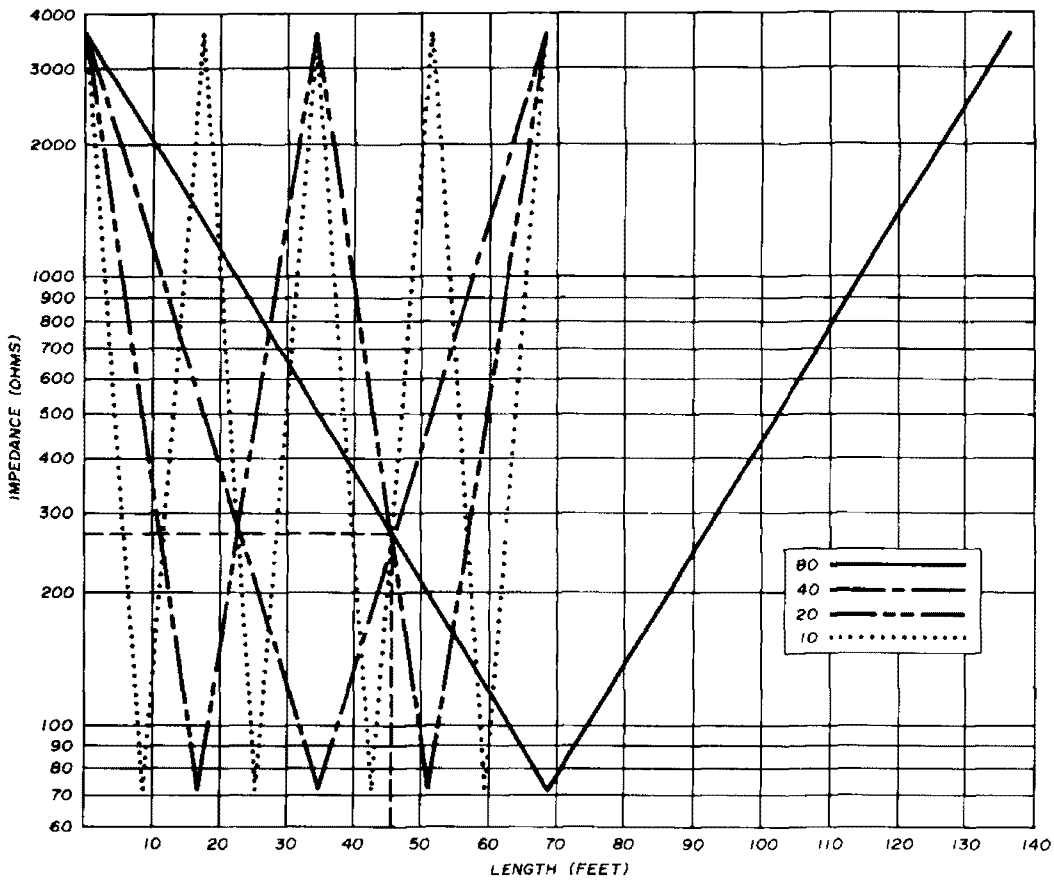


fig. 3. Impedance along a 136-foot antenna on 80, 40, 20 and 10 meters. Approximately 45 feet from one end of the antenna the impedance plots cross at 270 ohms — providing a fairly good match to 300-ohm feedline.

If the impedances present along this 136-foot antenna are plotted for the 80, 40, 20, and 10-meter bands, at a point 45 feet from one end, all four band plots cross at about 270 ohms (see **fig. 3**). If the antenna wire is broken at this point and the two wires are fed with 300-ohm twinlead, a fairly good match will be obtained for all four bands. In practice, certain lengths of feedline have been found to be preferred for easier transmitter loading. These lengths are multiples of 44 feet.

The advantages of both types of feedline, coax and twinlead, can be achieved by combining the optimum length of 44-feet of 300-ohm twinlead with a balun to match 75-ohm coax. A random length of coax can then be run to the hamshack as shown in **fig. 4**.

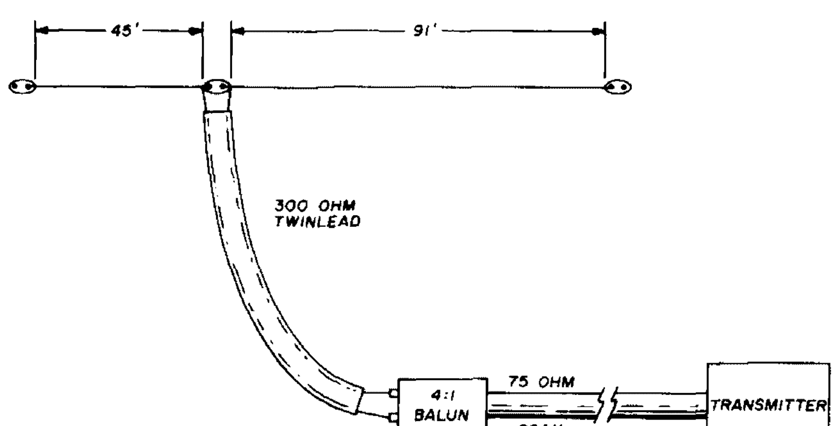


fig. 4. Windom antenna for four amateur bands uses 300-ohm twinlead, a 4:1 balun and 75-ohm coax to the transmitter.

construction

After obtaining 140-feet of number-12 Copperweld antenna wire, three egg insulators and 50-feet of 300-ohm twinlead, you are ready to proceed. Using the dimensions shown in fig. 4, install the three insulators. The distances shown are *between* insulators. Attach the premeasured 44-feet of 300-ohm twinlead (or multiples thereof) between the feedpoint insulator and the balun. Install the antenna as high and as in the clear as possible.

Route the 300-ohm feedline away from the feedpoint at a 90-degree angle for as far as possible. The balun should be waterproofed if it is exposed to the weather. One method is to completely wrap it with Scotch Brand vinyl tape of the type used by electricians and carried by most hardware stores.

There are several good commercial broadband baluns on the market that can be used, as well as toroidal kits for assembling a kilowatt unit in a small Minibox. The *ARRL Handbook* provides construction details for an easily made toroid balun.

There is one note of caution that applies to any multiband antenna system. Any harmonics generated on the lower bands will be efficiently radiated by this antenna. A conventional antenna tuner can be substituted for the balun, or used at the transmitter end of the coax to eliminate harmonics reaching the antenna. However, the use of an antenna tuner defeats the basic simplicity of the balun-to-coax feed system with its automatic bandchanging and no tuning to fuss with. Several excellent antenna tuners have been described in the amateur magazines.[1,2,3]

references

1. Ed Noll, W3FQJ, "Antenna Tuners," *ham radio*, December, 1972, page 58.
2. Gregory Widin, WB2ZSH, "Medium Power Toroidal Antenna Tuner," *ham radio*, January, 1974, page 58.
3. Ed Marriner, W6BLZ, "Match Box Antenna Tuner," *73*, September, 1966, page 38.

ham radio

the ham notebook

spurious causes

While intruder-watching, and doing associated Official Observer work when an unusual signal turns out to be amateur rather than an intruder, much is learned about the causes of troubles.

Recently, there have been noted many cases of key chirps or clicks spaced many kilohertz from the normal signal. Sometimes it is noise, which may be keyed normally or back-keyed, or voice peaks in phase with the desired signal. Several hams have found tubes to be the cause, though plugging the offending tube back in the same socket may not again give trouble.

It has been suggested that this may be the result of generation of spurious frequencies due to a temporarily corroded tube pin or socket contact, which is self-cleaned by removing the tube and plugging it in again.

Therefore, it is suggested that all tubes in transmitters and receivers, particularly those associated with the generation of the transmitted signal, periodically be wiggled or even pulled out and plugged back in. It would seem that this could be done several times a year, to keep contacts clean, so that some screen or suppressor does not lose its voltage or its rf ground connection.

Bill Conklin, K6KA

line voltage monitor

It should be of interest to most hams to know the deviation from normal line voltage available at any time in their shack. Several line voltage monitors have been described, but these generally have been complicated by incorporating features that are not necessarily required. Self-calibration, for instance, requires a significant increase in the number of components as well as requiring high-cost, precision items.

The expanded scale-line voltage monitor I have built reduces the number of components significantly and does not compromise the accuracy to any great extent. As indicated by the schematic in

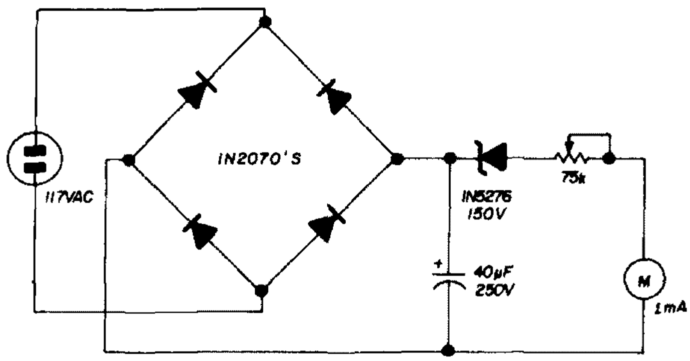


fig. 1. Simple expanded-scale line-voltage monitor reads from 115 to 125 volts on a 1-mA meter. Calibration is discussed in the text.

fig. 1, it consists of five active components, all of which were scrounged from my junk box.

However, for the recently licensed ham who may not have accumulated any sort of junk box, the cost of new parts, including the Minibox, cord terminal strips, etc., is under $7.50 (half of which goes for the milliammeter).

The whole circuit is mounted on terminal strips, so isolated from the chassis. Calibration was accomplished on a one-time basis using a Variac and a Simpson vom. With the Variac adjusted for 120 volts ac, the potentiometer was adjusted to give a mid-scale reading. The voltage was varied to 115 volts ac and 125 volts

ac and the meter indication went to either end of the scale.

Intermediate points at 1-volt intervals were marked on the face of the meter for instant reference. Although a trace of non-linearity was detected on the high side the line-voltage monitor tracked remarkedly well to within a few percent.

Finally, if it is desired to obtain a larger variation (i.e., plus or minus 10 volts or some other value) a lower value zener, 140 volts or lower, may be substituted.

Alfred J. Parker, WA8VFK

two-meter power amplifier

TRW Semiconductors has announced the first in a series of reasonably priced, npn power transistors designed specifically for amateur radio equipment. The first transistor in the series, the PT5757, provides 10-watts output at 150 MHz with a 12.5-volt power supply and is designed for operation on the 2-meter amateur band. A single PT5757 will boost the 1-watt output of a 2-meter rig to 10 watts. A simple circuit is shown in **fig. 2.**

The PT5757 can also be amplitude modulated to approximately 60% without damage. Ideal for mobile operation, the PT5757 has better than 70% collector efficiency at 10 watts and 12.5 volts. For hand-held rigs at reduced power, excellent performance can be obtained with collector voltages as low as 8 volts. Best of all, the PT5757 is priced at $10.00 in single quantities, and is available from any TRW distributor or from Ham Radio Center, 8342 Olive Boulevard, St. Louis, Missouri 63132.

Jim Fisk, W1DTY

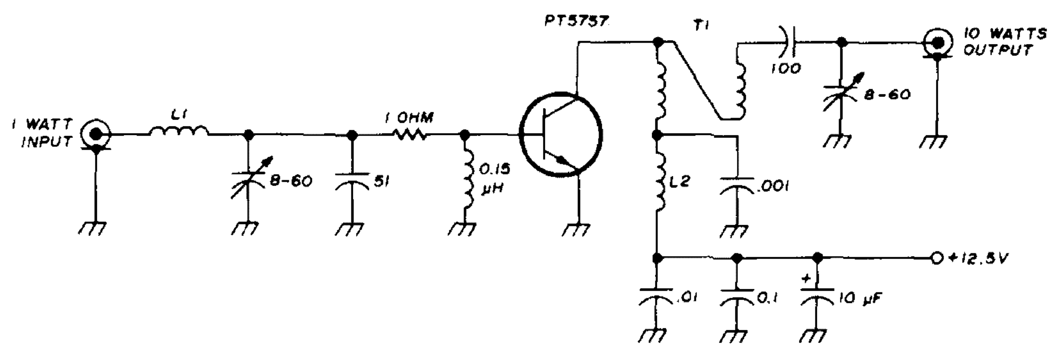


fig. 2. 10-watt 144-MHz power amplifier using the new TRW PT5757 transistor. L1 is 4 turns no. 20 enamelled, 3/32" ID; L2 is 10 turns, no. 20 enamelled, 3/32" ID. Transformer T1 is a 4:1 transmission-line transformer made from a 3" length of twisted pair, no. 20 enamelled wire.

ic lead former

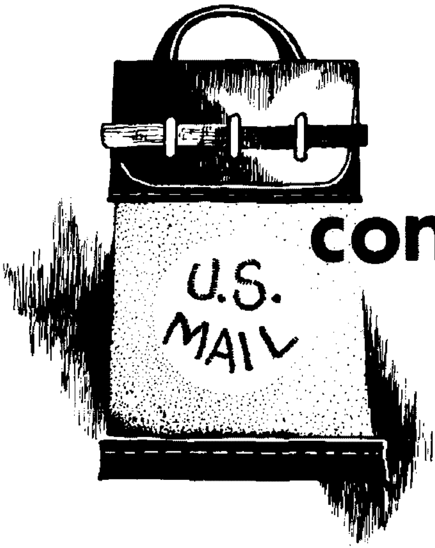
In making layouts for printed-circuit boards and in using breadboard circuits, difficulty has been encountered in connecting TO-5 can leads. Since the standard dual-inline-pack (DIP) configuration is very convenient for these applications, I decided to use this configuration for all ICs.

To accomplish the above, a lead former was constructed by drilling two rows of holes, 0.3 inch apart, with holes spaced on 0.1 inch centers in a piece of scrap printed circuit board stock. To use the lead former, the IC leads are inserted in appropriate holes, the IC pressed down, and the leads trimmed on the reverse side of the former.

In some applications, it is more convenient to use alternate holes (0.2 inch spacing) to provide additional spacing for an 8-lead IC such as the CA3028A. With this spacing, an 8-lead IC fits a standard 14-pin DIP socket. For further simplification, the unused leads of the IC may be clipped off near the can before connection.

This simple device provides a means of forming TO-5 can leads for the experimenter, simplifying his layouts.

Bill Stauffer, W5ICV



comments

receiver selectivity

Dear HR:

There have been articles over the years, and several recently, outlining the advantages and desirability of improving the front-end selectivity of the receiver. To see how a receiver of recent design checked in this respect, I checked out a Hammarlund HQ-215. This is a solid-state version of the Collins 75S receiver. Both have a 200-kHz bandpass i-f between the first and second mixers. The HQ-215 has

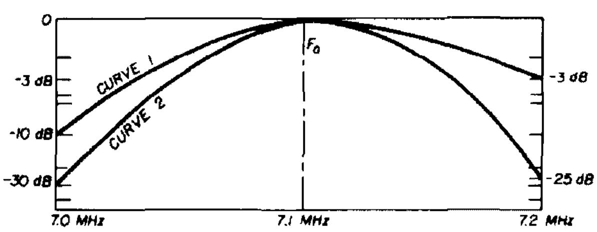


fig. 1. Passband characteristics of the HQ-215. See K4ZZV's letter for operating conditions for curves 1 and 2.

three tuned circuits ahead of the first mixer – the 75S only two.

It can be seen from **fig. 1** (curve 1) that there is quite a wide passband to the incoming signals as far back as the second mixer input where the signal was recorded without agc. This was taken with the original rf swamping resistors removed to help the selectivity as much as possible!

Curve 2 was run with the i-f swamping resistors removed and everything peaked up for the band center. There are four tuned circuits in the HQ-215's first i-f. Nothing was changed in their intercoupling. Quite an improvement in the selectivity can be seen.

The noise figure of the receiver was halved with the additional gain so the first rf tuned circuit was stepped down enough to restore the original noise figure which would also improve the front-end overload characteristics. It looks like the next project should be a vari-cap network ganged to the vfo to make full use of the improved first i-f selectivity.

There are receivers that do tune the first i-f to their advantage. The Collins 75A4 is a good example. Getting as much selectivity as close to the antenna as possible really makes for the ultimate in reception. If one wants to confine their operation to a few Hz as the fixed-channel stations do, or guard a special traffic net frequency, a crystal filter at the antenna input is just the thing to clean things up. Maybe some day a variable frequency filter will be developed that will do what a few coupled LC circuits cannot.

There are trade-offs in receiver design though, and at least one solid-state receiver uses a tube in the front-end to handle the strong signals. It should make the old timers happy to know that tubes are still being used in modern design.

Wayne W. Cooper, K4ZZV
Miami Shores, Florida

code speed

Dear HR:

I certainly enjoyed VE2ZK's recent article on code speed, which mentioned that the FCC and ARRL use 50-bit words such as PARIS to establish the speed, and other government agencies use the 60-bit word CODEZ.

As a series of dots have equal "up-key" spaces between them, a string of dots is actually a number of dot cycles. Thus 25 dot cycles is the same as 50 continuous bits. This might clarify the author's mention of dividing by 25 to get the speed.

However, one of the most simple methods of determining speed without any mathematics is to merely send continuous number one (1) characters. Count the number sent in 24 seconds and this is the wpm for 50-bit rate. For 60-bit rate, count the 1's sent in 20 seconds. You could also use a digital counter with an automatic keyer although some counters might give erroneous results on the pulsed output of the keyer. Divide the counter reading by 25 to indicate wpm. It should at least get you in the ballpark, and many amateurs own digital counters these days.

The article was very interesting and it is the first time I can recall any author attempting to explain the 50- and 60-bit words and where they are used.

Irv Hoff, W6FFC
Los Altos Hills, California

attenuation pads

Dear HR:

I found the comments by Mike Goldstein, VE3GFN, *on tuner* overload memory in *ham notebook* of the January, 1973, issue, provoking. The problem I had was with converter overload when I was operating on 6 meters. Two other hams in my town who operated on 6 meters lived within half a block of me, so the old tubes of the converter would really light up when either of them came on.

I wanted to attenuate incoming signals *without changing* the impedance of my receiving system. A T-pad is just the device to do that. The Mallory RT-50, a 50-ohm pad, while designed for audio work, performed beautifully for me. As an experiment, I put it between my Heath Mohawk and the International Crystal converter and found that with the attenuation control set to zero, signals came in stronger on the Mohawk with the T-pad in the circuit than without it. I attribute this to better impedance matching created by inserting the T-pad between the converter and receiver. The pad would not only be helpful for converter overload due to strong signals, but also to receiver overload caused by too much converter output.

The Mallory RT-50 pad comes with knob, dial plate, mounting hardware and hook-up instructions, and can be obtained from many electronic stores and mail-order houses. If you have difficulty obtaining one, you can send $3.60 plus postage to Scott Electronic Supply Corporation, 4040 Adams Street, Lincoln, Nebraska 68504.

James Worrest, KØHNQ
Lincoln, Nebraska

sporadic-E openings

Dear HR:

The article on predicting sporadic-E openings by Morrie Goldman in your October, 1972, issue is quite informative. In fact, its usefulness extends beyond the author's original purpose. Several times in the past I have been plagued with spurious responses in my receiving equipment which were caused by the presence of nearby high-power paging transmitters. Your **table 1** will be quite useful in chasing down these problems in the future.

A second point which radio amateurs should find useful is the direct correspondence between **table 2** and our amateur call areas.

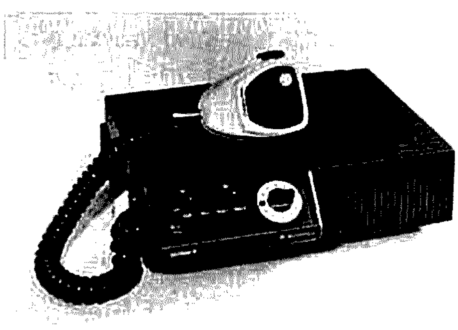
WØ, KØ	KA, KB	W5, K5	KK, KL
W1, K1	KC, KD	W6, K6	KM, KN
W2, K2	KE, KF	W7, K7	KO, KP
W3, K3	KG, KH	W8, K8	KQ, KR
W4, K4	KI, KJ	W9, K9	KS, KT

This correspondence makes it unnecessary to have to continually refer to the chart while monitoring a band opening.

Lewis D. Collins, K4GGI
Arlington, Massachusetts

new products

motorola vhf-fm radio for amateurs



Motorola, long a leader in two-way vhf-fm equipment, has now entered the amateur vhf-fm market through their subsidiary, Modar Electronics Inc., with the introduction of the new *Metrum II* two-meter vhf-fm radio. This radio, which covers the 144-148 MHz amateur band, is totally solid-state unit with a number of unique features. It is offered in 10- and 25-watt versions, both switchable to 1 watt.

The modern, attractively styled line features a shadow bronze finish, 12-channel capability, the dependable Motorola microphone and field-proved circuitry. It incorporates a rotary on/off volume control, a variable squelch control, an illuminated control instrument panel, detent high-low power, detent repeater input and two detent auxiliary switches for custom adaption by the radio operators.

With built-in antenna mismatch protection, the *Metrum II* radio will continue functioning without damage to the unit even if the antenna is damaged, disconnected or improperly connected. Reverse polarity protection provides added safeguards against improper installation.

A specially designed reversible control panel allows the radio to be mounted in almost any position while maintaining clear visibility of all controls. A universal mounting tray permits installation at virtually any location. Indirect, non-glare back lighting of the *Metrum II* control panel means all controls can be read easily. Optional accessories include ac power supply, quarter-wave whip antenna, crystals, dimmer mod kit and rf indicator kit.

Manufacturer's suggested list prices for the 25-watt, 12-channel model and the 10-watt, 12-channel model version are set at $499.95 and $399.95, respectively. For further information on the *Metrum II* fm amateur radio, write to Modar Electronics, Inc., 2100 North Meacham Road, Schaumburg, Illinois 60172, or use *check-off* on page 110.

fm modulation meter

The ECM Corporation has announced the first commercially available fm modulation meter designed especially for the amateur. The ECM-5 covers all ham bands between 52 and 450 MHz, and features a peak reading meter. Deviation of any fm transmitter can be accurately adjusted between 5 kHz and 25 kHz in seconds, using voice or tone modulation.

More Details? CHECK–OFF Page 110

The ECM-5 fm modulation meter closely follows the circuits used in professional equipment except that frequency is crystal controlled. This allowed ECM engineers to eliminate many expensive circuits needed only when frequency selection is vfo controlled. The net result was a tremendous reduction in price without sacrificing quality. The frequency selecting crystals are the popular, subminiature, third-overtone type used in many of today's fm receivers. These crystals were chosen for their low price and availability.

The peak reading meter has a special time-constant circuit that causes the needle to deflect upscale rapidly and downscale slowly. This allows the needle to follow voice peaks and increases the accuracy of readings when checking deviation using voice modulation. Other features include a builtin or external antenna, all solid-state construction, battery powered by inexpensive AA pencells, and a battery condition indicator.

The ECM-5 is priced at a low $75.00, less batteries and crystals. For more information, write ECM Corporation, 412 North Weinbach Avenue, Evansville, Indiana 47711, or use *check-off* on page 110.

multifrequency antennas

An antenna farm in your own backyard! That is what Don McVicar, VP7DX/VE2WW, has claimed to have developed in his new Mark IV, V and VIII multifrequency directional wire beams. Don has been experimenting many years with antennas and has developed an allband antenna system which is economical, mechanically sound, inconspicuous and easy to install. It gives good gains with a low angle of radiation at moderate installation heights.

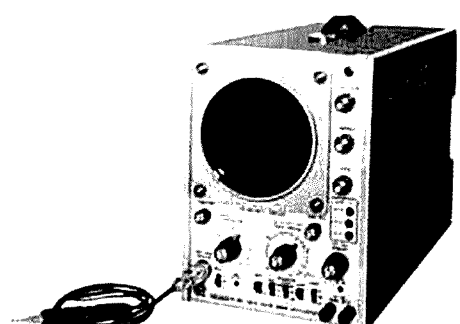
Electrically, the Mark IV antenna has a minimum theoretical forward gain over a reference dipole of about 5.5 dB. While this is encouraging, consistent forward gains of up to 3 S-units have been achieved on both transmission and recep-

tion. Don assumes that these low-angle gains are due to the sloping inverted-vee configuration. On 40 and 80 meters the front-to-back and front-to-side ratios are from 9 to 20 dB and deep pattern nulls are not evident. On 20, 15 and 10 meters the front-to-back and front-to-side ratios vary between 12 and 30 dB. The feed-point impedance is 52 ohms, power capacity is 2-kW PEP and the antenna uses a single-line, all-band feed system.

Mechanically the system consists of two or more sloping inverted-vee antennas, one-half or 3½-wavelengths long, fed at their apex through a unique rf switch which causes the elements to react parasitically on one another. The switch is built for heavy-duty outdoor use. The antenna wire (not supplied) can be number-14 or equivalent, and the system tends to be self guying.

The rf switch may be installed on an existing rotator shaft beneath a Yagi or quad on any mast or tower such as an inexpensive 50-foot telescoping mast. For more information on this unique antenna system, write to World Wide Antennas, Box 467, Miami Springs, Florida 33166, or use *check-off* on page 110.

triggered-sweep 10 MHz oscilloscope



The new Eico Model TR-410 oscilloscope claims to be the industry's lowest priced lab quality, wideband triggered-sweep oscilloscope. It is expressly designed for speedy precision servicing, lab work, production testing and vocational instruction with such advanced features as automatic sweep which locks with complex tv signals, 10-MHz bandwidth,

More Details? CHECK—OFF Page 110

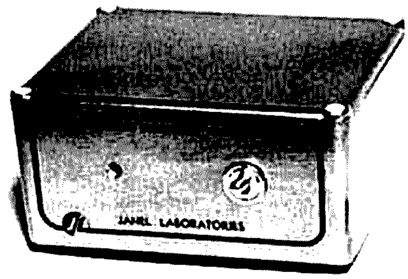
all solid-state design with protected fet input stage, and single dual probe to convert quickly from direct to 10:1 low capacitance operation. The instrument may be operated from a standard 120-volt line, low 100-volt or 220-230 volts – all 50/60 Hz. Included are three calibration voltages, 2, 5 and 10. The horizontal and vertical dc balance controls are adjustable with a screwdriver from the front panel for convenience and accuracy. Included are vertical and horizontal selection of ac or dc modes of amplification. The gate signal is available at a jack to enable the operator to synchronize other equipment to the trace displayed on scope. The astigmatism control is on the rear panel because once it has been set, readjustment is seldom, if ever, required. The removable sides, top and bottom provide easier and more accurate servicing and calibration. Standard bezel and bushings are provided for camera mounting.

The new Eico model TR-410 oscilloscope is priced at $379.95 and is available from your local Eico dealer. For more information, use *check-off* on page 110.

voltage-controlled attenuators

An economical series of three voltage controlled PIN diode attenuators cover the frequency range of 5-200 MHz for agc or leveling or other closed loop applications. Models VCA-1, 40 dB, 5-100 MHz; VCA-2, 30 dB, 5-100 MHz; VCA-3, 20 dB, 5-200 MHz; are offered. Maximum insertion loss is 6 dB. Vswr varies from 3.0 to less than 1.5:1, depending upon the attenuation setting. Rise and fall times of attenuation to specified values permit wideband modulation of rf signals. Units require up to 105 mA supply current and less than 5 mA control current. Positive or negative supply and control voltages may be specified. Connectors available include BNC, JCM (SMA compatible) and TNC. For more information, use *check-off* on page 110, or write to Radiation Devices Company, Post Box 8450, Baltimore, Maryland 21234.

low-noise 432-MHz preamps



As part of an expanding line of quality vhf/uhf products, Janel Labs has announced a series of 432-MHz preamps. Four models are available, offering low noise figures in a choice of two price ranges, each having the option of an ac power supply. Models without power supply (432PA and 432PC) have a compact sheet aluminum enclosure while those with power supply (indicated by the suffix −1) feature a rugged cast-aluminum case.

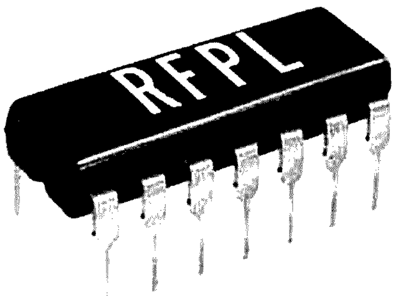
The gain of all models is an ample 20 dB. The 3-dB bandwidth is about 20 MHz. Stock units can be supplied for any center frequency between 420 and 470 MHz. Other frequencies are available on special order.

The basic circuit is a two-stage amplifier. This uses a KMC bipolar transistor first stage and a 3N159 dual-gate mosfet second stage. The 432PA uses a K2073 first stage to produce an outstandingly sensitive 3.5-dB noise figure. The 432PC uses the new K6007 to achieve an extremely sensitive 1.5- to 2.0-dB noise figure. The low cost 432PA and 432PA-1 are expected to see wide use for 450-MHz fm as well as for general purpose applications such as DX, ATV, and OSCAR.

The 432PC and 432PC-1 meet the needs of the most demanding applications such as moonbounce and weak-signal CW work. Prices range from $29.95 for 432PA to $94.95 for the 432PC-1. All are postpaid and guaranteed. For more information, write to Janel Laboratories, Box 112, Succasunna, New Jersey 07876, or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

rf directional couplers



RF Power Labs have introduced a new line of low-cost miniature wideband rf bidirectional couplers which should be very useful to amateurs who build their own high-frequency and vhf equipment. These versatile couplers can be used for power sampling for waveform monitoring and power level checkpoints, for load impedance and vswr measurements, and for direct readout of forward and reflected power over wide frequency ranges with excellent accuracy. Units are available in both dual-inline packages and *flat pak* configurations, and are capable of handling rf power levels up to 3 watts over their specified bandwidth.

Four models of the bi-directional coupler are available: the DC-14/14A, covering 2 to 300 MHz; the DC-14B/14C, covering 1 to 300 MHz; the DC-14D, covering 500 kHz to 100 MHz; and the DC-14E, covering 50 kHz to 100 MHz. Prices in small quantities range from $13.90 to $15.90 each. For more information, write to R.F. Power Labs, Inc., 92 - 104th Ave. N.E., Suite 103, Bellevue, Washington 98004, or use *check-off* on page 110.

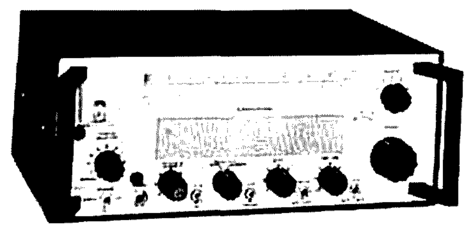
digital catalog

ES Enterprises has announced the availability of a new 6-page catalog that describes their total line of digital products. Standard products include low cost programming instruments and controls, timers, clocks, counting and measuring devices. Also included is a complete listing of their modular display units for custom digital instrumentation

and a section outlining their digital clock, multimeter and frequency counter kits.

The catalog contains descriptions of basic features, custom capability, standard options and warranty information. A price list and ordering instructions accompany the catalog. For a copy of the catalog, write to ES Enterprises, 10418 La Cienega Boulevard, Inglewood, California 90304, or use *check-off* on page 110.

general-coverage receiver



Up to ten crystal-tuned frequencies can be preselected for drift-free automatic tuning on the latest high-quality general-purpose communications receiver from the British firm of Eddystone. The Model 1001 receiver has the unusual feature of a rechargeable internal power supply, consisting of a nickel-cadmium cell, which serves as a temporary standby in case of main circuit failure. The set will also work off an external 12V dc battery supply.

Covering frequencies from 550 kHz to 30 MHz, the set incorporates automatic gain control, wide/narrow selectivity control, carrier-level meter and both coarse and fine tuning knobs.

Frequency calibrations for tuning are marked on a horizontal scale drum which splits the 550-kHz to 30-MHz range into five bands. The scale display is about 6-1/2" long; a secondary scale below the main calibration can be used in conjunction with a vernier dial to provide a useful logging facility. Dial illumination can be switched on and off.

Aimed at the serious radio amateur, short-wave listener and DXer, the Model 1001 is designed to the most stringent

More Details? CHECK—OFF Page 110

professional specifications, and incorporates a variety of solid-state devices including integrated circuits and field-effect transistors. It drives its own pair of miniature speakers, and has output facilities for headphones, external speaker and tape-recorder. Price of the Model 1001 is about $900 delivered, including duty and taxes. The North American distributor will welcome inquiries from U.S. customers and prospective dealers. Write to Conway Electronic Enterprises Ltd., (Mr. J.W. Cave, General Sales Manager) 88/90 Arrow Road, Weston, Ontario, Canada, or use *check-off* on page 110.

radio transmitter principles and projects

Amateur radio operators, communications technicians and transmitter experimenters will profit from this new and completely up-to-date book by Ed Noll, W3FQJ. Devoted entirely to the subject of radio transmitters, this book also is perfect for those studying for the various grades of amateur or commercial FCC license examinations.

The first three chapters contain information on electron devices — the fet, bipolar transistor and the vacuum tube. Different modes of modulation — CW, a-m, fm and ssb — are discussed in other sections. Chapter 4 describes hybrid transmitter circuits using tubes and transistors. Double-sideband and single-sideband generation and circuits are covered in Chapter 5. There is a chapter on linear amplifiers and mixers; another explains integrated circuits. The final three chapters cover vhf circuits, frequency modulation and transmitter testing.

Each chapter begins with basic principles and advances to more detailed information. The projects are based on the basic principles and are designed to further the reader's understanding through actual experience. They also provide the radio amateur with complete plans to build his own gear. 320 pages, $6.95 (softbound). Order from Comtec Books, Greenville, New Hampshire 03048.

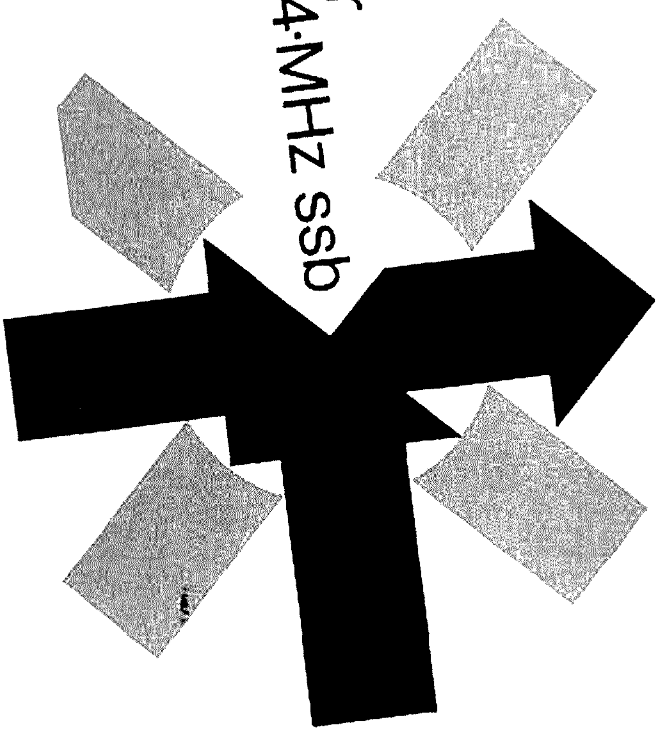
85 cents

focus on communications technology . . .

ham radio magazine

FEBRUARY 1974

solid-state TRANSMITTING CONVERTER for 144-MHz ssb



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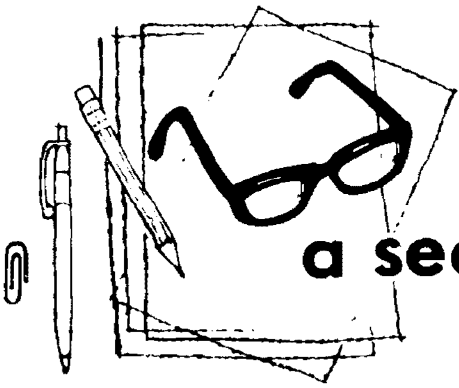


a second look

by jim fisk

The first two issues of *HR Report* are now off the presses and in the hands of subscribers around the country. If you want to know what's happening behind the scenes in amateur radio, and rapidly, as the news breaks, *HR Report* is the only way to do it. For example, did you know that we will probably lose the upper 2 MHz of the 420-MHz band (448 to 450 MHz) to the Emergency Medical Service? Did you know that the ARRL's first ten-meter contest was a partial success, with openings to Africa and South Pacific? Did you know that more than 500 two-meter repeater licenses have been issued by the FCC, nearly clearing up the backlog? Did you know that a large variety of quality-made coils, chokes and terminal boards, packaged for the amateur, are now available from Cambridge Thermionics Corporation (CTC)? These are just some of the items covered in detail in recent issues of *HR Report.* For subscription details for this new bi-weekly amateur newsletter, look on page 72.

This month we will kick off the latest project of our *more for 74* program, an Automatic License Renewal Service for *all* FCC-licensed amateurs (except Novices), subscribers to *ham radio* or not. The cost to you? Absolutely nothing, except for the effort to open an envelope. It will work like this: 60 to 90 days before your amateur license is due to expire you will receive in the mail a copy of FCC Form 610 plus a supplementary instruction sheet prepared by our staff which will include some info on such things as renewal fees, operating after your license expires *if* you filed a timely renewal application, etc. All you have to do is fill out the Form 610, enclose your check or money order, and mail it back to the FCC.

No last-minute scurrying around to find the proper form, or discovering that your license expired a month ago. However, for the Automatic License Renewal Service to operate successfully, your correct address must be on file with the FCC as required by the regulations. If you have moved since you last renewed your amateur license and the FCC does not have your current address you will not receive your License Renewal packet. It's as simple as that.

Early this month the first Automatic License Renewal packets will be in the mail to amateurs whose licenses expire in March and April, 1974. In early March License Renewal packets will be mailed to amateurs whose licenses expire in May. From then on mailings will be made the first of every month so you should receive yours at least 60 days before your license expires.

Jim Fisk, W1DTY
editor-in-chief

solid-state transmitting converter for 144-MHz ssb

Complete construction details for a solid-state transmitting converter that provides more than 30-watts output on two meters

Bob Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022

In the past few years numerous articles have been published describing transistorized fm and CW transmitters and class-C power amplifiers for use in the 144- to 148-MHz band. However, there has been a noticeable lack of information covering single-sideband applications or *linear* transistor amplifiers, due primarily to the problem of generating reasonable amounts of power in linear amplifiers operating in the vhf region.

When my faithful but venerable tube-type two-meter ssb transmitter started to show its age, I began to investigate the possibilities of replacing it with a solid-state unit. A review of the published literature and manufacturers' data indicated that there should be no major obstacles up to the 1-watt level, but I found little encouragement to attempt the 30- to 35-watt output I was seeking. Fortunately, I had available a substantial quantity of vhf power transistors, designed for class-C service, with which to experiment. This article will show that it is entirely feasible to operate vhf power transistors as linear amplifiers, using techniques which are well within the capabilities of the serious vhf experimenter.

prerequisites

My needs were to generate a minimum of 30 watts PEP between 144 and 146 MHz, using my high-frequency ssb transmitter as the basic exciter. This dictated using the 28- to 30-MHz output from the transmitter in order to cover a 2-MHz range without changing the local-oscillator frequency in the transmitting converter. However, a close examination of the mixing scheme revealed one dismaying problem—operation at 145 MHz requires the ssb input to be at 29 MHz, and the fifth harmonic of 29 MHz is also 145 MHz, which is most undesirable. Since most mixers are excellent harmonic generators, I had to find one that was not. Luckily, the double-balanced mixer

has excellent characteristics in this respect, thus eliminating one stumbling block.

I also decided that if I were going to have major problems (more than just the expected ones), there was no point in constructing the entire unit, and that if a problem was to prove insurmountable, it would show up before the final stage. Therefore the logical approach would be to build up the circuit to the driver stage, and then cover the final amplifier as a separate subproject. As it turned out, this

values may be found in reference 1 or any standard reference text.

I used a double-balanced mixer board which had been given to me because of a broken wiring trace. Its characteristics are identical to several inexpensive mixers now available, such as the Anzac MD108, Mini-Circuits SRA1, and Vari-L DBM166. Any one of these will be suitable, as would be the more expensive Hewlett-Packard or Relcom models. I have recommended the Anzac mixer in the parts list for fig. 2, since it is the least costly

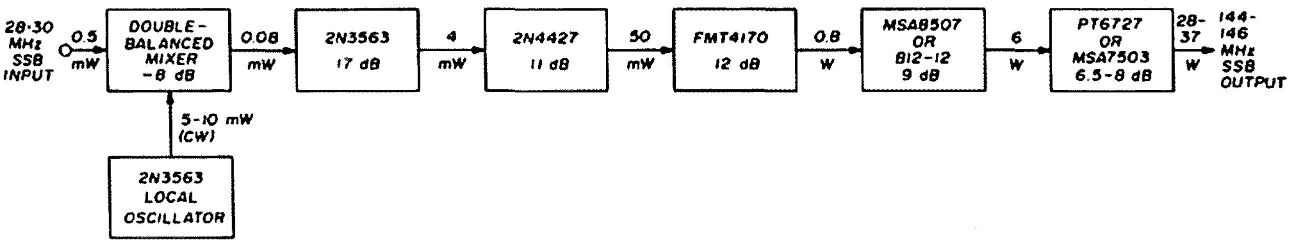


fig. 1. Block diagram showing approximate stage gains and peak-envelope-power levels throughout the converter. Output power and gain of the final stage depend on the type of transistor used and the collector supply voltage.

was a fortunate decision, since it allowed me considerable flexibility in designing the final stage.

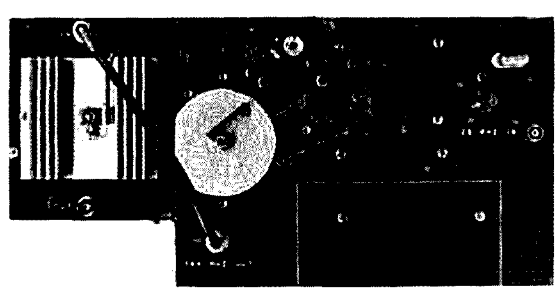
driver unit

A complete block diagram of the transmitting converter appears in **fig. 1**, showing the *approximate* stage gains and power levels throughout the circuit. The schematic of the driver unit is shown in fig. 2. The 28- to 30-MHz ssb input is applied to the RF (*R*) input port of double-balanced mixer Z1, and should not exceed -3 dBm (0.5 mW) to keep distortion products to a minimum. The 50-ohm pad formed by resistors R1 through R3 has been included to insure proper termination of the transmission line from the SSB exciter and to provide the mixer with a 50-ohm source. Values have not been specified for the pad resistors, since the required attenuation will depend on the output power of the exciter and the amount of attenuation present in your external power attenuator. The total loss in the two attenuators must be sufficient to limit the input to the mixer to the specified 0.5 mW. Design equations for calculating the resistance

and is directly available from the manufacturer in single-lot orders.

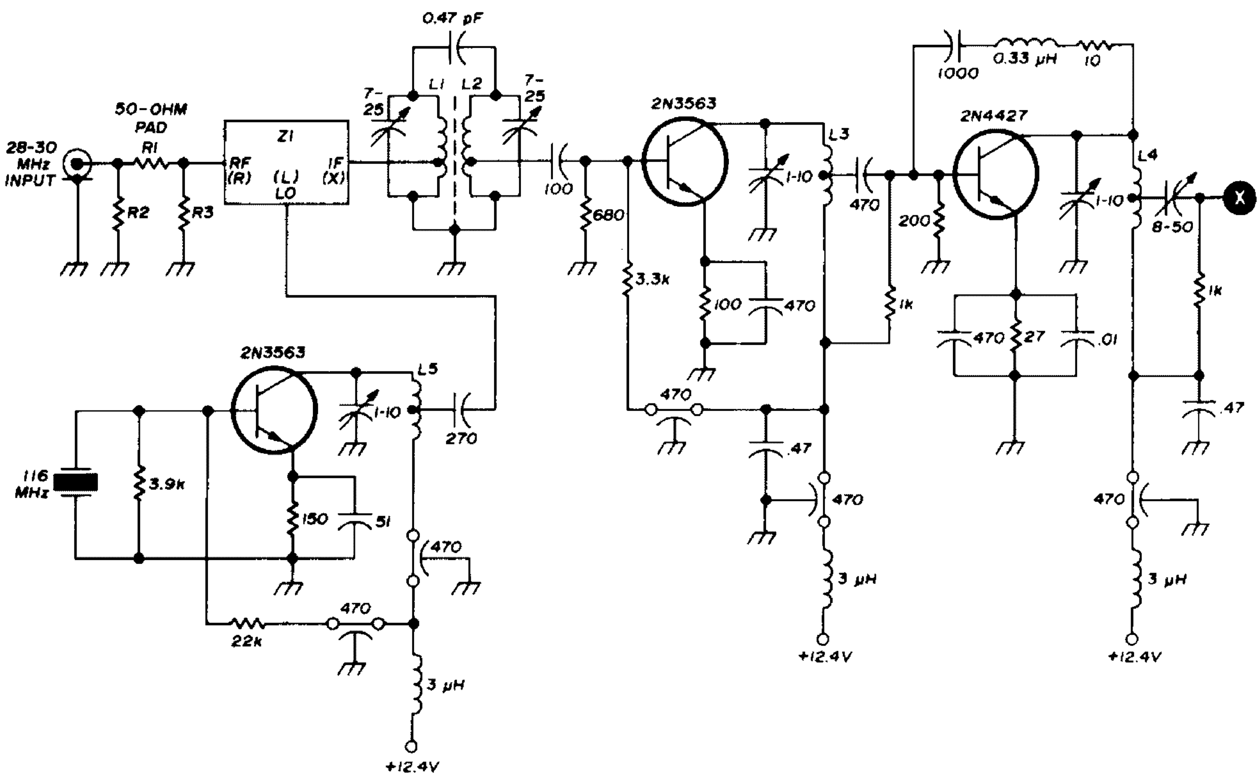
In order to achieve minimum loss through the mixer, a local-oscillator signal of at least 7 dBm (5 mW) is required. This is easily obtained from a 2N3563 operating in a Miller oscillator circuit which uses a 116-MHz overtone crystal. The oscillator output is taken from a tap on the collector coil, chosen to provide maximum power transfer to the mixer.

The output of the mixer is obtained at



Top view of the driver unit and MSA7503 final amplifier. The metal disc is the heat sink for the driver transistor; top-hat heat sinks are used on the 2N4427 and FMT4170 transistors. The 1N4001 and 1N4719 diodes are inside the clamps on the driver and amplifier mounting studs, respectively.

the i-f *(X)* port and is applied to a double-tuned top-coupled filter, resonant at 145 MHz. The Q of each tuned circuit and the coupling coefficient have been selected for a bandwidth of approximately 4 MHz. The input and output taps on L1 and L2 provide impedance matching additional .01-μF emitter bypass capacitor were included to suppress a tendency of this stage to oscillate. The output of the 2N4427 is matched to the base of transistor Q1 by means of a trimmer capacitor tapped down on the collector coil.



C1,C3	1.5-20 pF compression trimmer (Arco/El Menco 402)	L6	copper strap, 1¼" long x 3/16" wide
C2,C4	7-100 pF compression trimmer (Arco/El Menco 423)	L7	3 turns no. 16, ¼" ID, 5/8" long
L1,L2	4 turns no. 20, ¼" ID, ½" long, tapped ½ turn from ground end	L8	*25 turns no. 26E wound on Micrometals T37-3 core
L3	5 turns no. 20, ¼" ID, 3/8" long, tapped ½ turn from supply end	Q1	*Fairchild FMT4170, RCA 2N5913, or Motorola HEP-S3001
L4	6 turns no. 20, ¼" ID, ½" long, tapped 1 turn from supply end	Q2	*Fairchild MSA8507 or CTC B12-12
L5	6 turns no. 20, ¼" ID, ½" long, tapped ½ turn from supply end	R1,R2,R3	see text
		Z1	*double-balanced mixer (Anzac MD108 or equal)

fig. 2. Schematic diagram of the driver unit. Sources of asterisked items are listed in the appendix. All 1-10 pF capacitors are piston type.

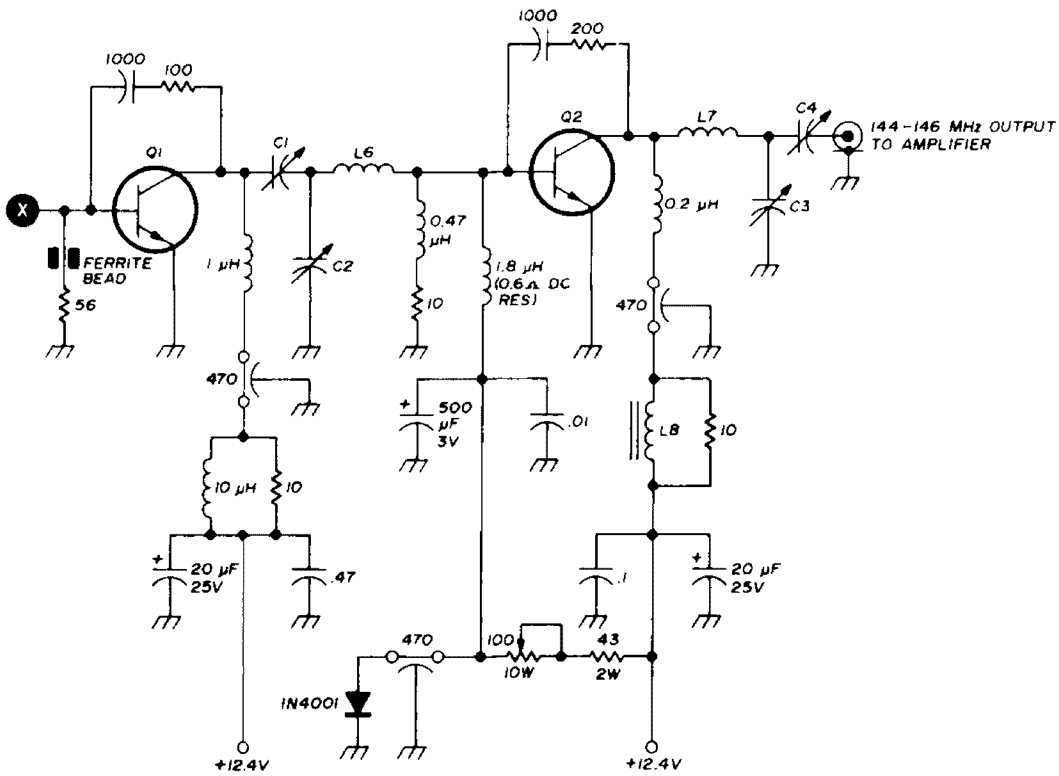
into and out of the filter. Total loss through the mixer and filter is about 8 dB.

Another 2N3563 follows the mixer, configured as a conventional class-A amplifier. The output of the 2N3563 drives a 2N4427 amplifier, which also operates class A. The RLC network between the collector and base of the 2N4427 and the

Several types of transistors were tried at Q1, all with some degree of success. The best of these was found to be the Fairchild FMT4170, although the lower-priced 2N5913 or HEP-S3001 (in that order of preference) should also be satisfactory. This stage operates closer to class-AB than class-A to keep the transistor power dissipation within acceptable

limits. The RC network between the collector and base improves the linearity of the stage. Parasitic oscillations in the hf region are suppressed by the 10-μH rf choke in parallel with a 10-ohm resistor, plus the large bypass capacitors in the collector supply circuit.

To prevent thermal runaway of the transistor, the base current is controlled by a 1N4001 diode which is thermally coupled to the transistor case. This is physically accomplished by mounting the diode on the stud of the transistor so that it follows the temperature of the device.



The driver stage was designed around a Fairchild MSA8507 vhf power transistor which is characterized only for class-C operation. The transistor is forward biased into class AB operation by means of a bias circuit described by Roy Hejhall, K7QWR.[2] Quiescent collector current is set by adjusting the base bias by means of the 100-ohm adjustable resistor.

It is essential that there be approximately one-half ohm dc resistance between the base and bias source for the bias circuit to operate properly. I used a 1.8-μH rf choke from my junk box because it had a resistance of 0.6 ohm. Any choke having an inductance between 0.47 and 2 μH will be satisfactory, provided that it has the required resistance. Otherwise a resistor may be inserted between the rf choke and the bias source to make the total resistance about 0.5 ohm.

Therefore, as an increase in transistor temperature tends to increase the base and collector currents, the increase in diode temperature causes the base bias to decrease, thereby reducing the base and collector currents to the equilibrium values set by the bias-adjust resistor.

The collector-to-base RC linearizing network and the collector-supply hf

table 1. Characteristics of the Fairchild MSA8507 transistor at 175 MHz with 12-volt collector supply.

P_{out}	12 watts minimum
P_{in}	3.5 watts maximum (at rated P_{out})
Z_{in}	1.5 + j1.3 ohms
Z_{out}	3 − j2.7 ohms
C_{cb}	35 pF (at 1 MHz)
BV_{CES}	36 volts
V_{CEO}	18 volts
I_c	2.0 amperes maximum
P_D	22 watts

parasitic-suppression network are both similar to those used in the preceding stage. Power is coupled into and out of the transistor by means of conventional T-networks, resulting in an output from this stage of approximately 6 watts PEP when fed into a 50-ohm load.

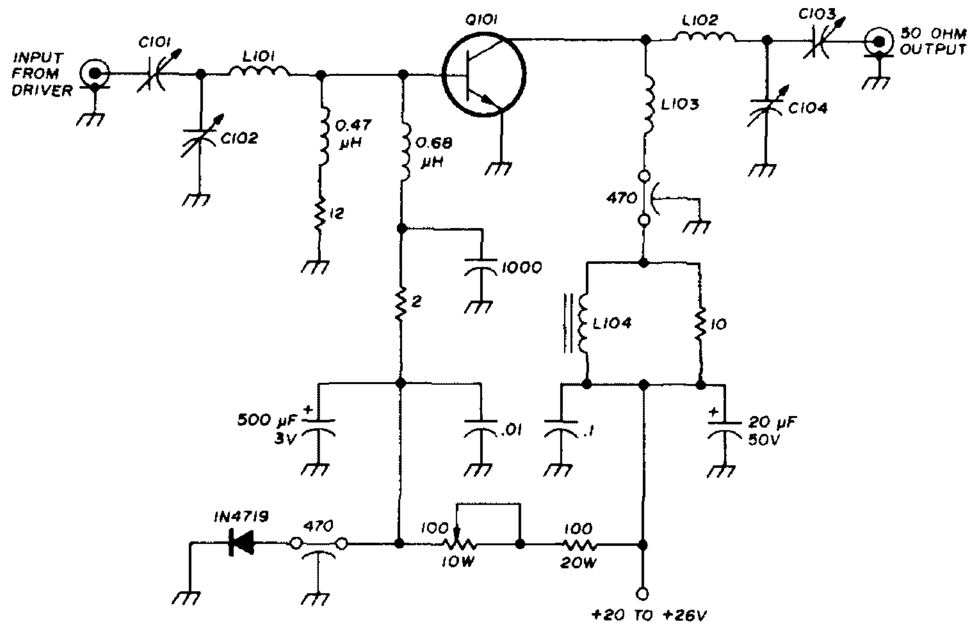


fig. 3. Schematic diagram of the final amplifier which uses a high-conductance diode to control base biasing. Details of the parts identified by reference designators appear in table 2.

Unfortunately, the Fairchild MSA8507 is no longer in production, although there may be some to be found as old stock or at surplus outlets. However the B12-12, manufactured by Communications Transistor Corporation, has similar characteristics and should be as good, if not better. For those interested in trying other transistors, the pertinent characteristics of the MSA8507 are listed in table 1. Reference 3 contains the design equations for the input and output networks, which must be redesigned if you use a transistor having input and output impedances substantially different from the MSA8507 or B12-12.

final amplifier

Two different amplifiers were designed, built, and operated on the air. The first uses a transistor characterized for class-C service in the 100- to 175-MHz region and having internal emitter stabilizing resistors. The circuit is shown in fig. 3. The first obvious question is why a nominal 28-volt transistor was used when the rest of the converter uses 12-volt devices. The answer is equally obvious when you look at the typical characteristics of 12-volt, 50-watt transistors—they do not have the necessary power gain. And as will be seen later, the dual voltage requirement is not a major problem.

The circuit configuration is similar to that used in the driver stage, and uses a Fairchild MSA7503 transistor. The input and output networks are designed to match a 50-ohm source and load, respectively. As with the MSA8507, the MSA7503 is also out of production. However, the technique of placing a 50-watt transistor in linear service, when it was designed for class-C operation, may be of interest. The bias circuit is the same as previously described for the driver stage, except for one minor difference. Because of the relatively large value of base current, an rf choke having less inductance but using larger wire was used in the base circuit. Therefore a 2-ohm resistor was added between the choke and bias source to provide an empirically determined optimum value of resistance.

The transistor operates as close to true class B as possible. That is, the base is just barely forward biased, so that the quiescent collector current is 2 or 3 mA. Considering that the peak dc collector current is about 2 *amperes,* that is truly class B. All attempts to increase the static collector current resulted in catastrophic failure of the device when excitation was applied, probably caused by secondary breakdown. (See reference 4 for a discussion of this phenomenon.) However, as long as the static collector current is limited to 3 mA or less, the amplifier is stable, reliable, and entirely satisfactory. The output powers obtained at collector voltages between 20 and 26 volts are shown in **fig. 4**.

A second amplifier was then designed and built, using a commercially available transistor and a different biasing scheme. A TRW PT6727 is used in the circuit shown in **fig. 5**. This transistor is emitter-ballasted and is designed not only for CW operation at 150 MHz, but for a-m service as well.

The heart of the bias network in this circuit is a device called a *byistor,* which is manufactured by Communications Transistor Corporation, and shown in **fig. 5** as a Y-shaped symbol (originated by CTC) with its type designation BY1. The byistor acts as a low-impedance dc bias source and consists of a diode and silicon resistor; **fig. 6** shows the internal arrangement. The device is packaged in a ceramic stripline configuration, identical to that used for rf power transistors, and is meant to be mounted on the same heat sink as the transistor for temperature tracking. The diode is fabricated using the same material, geometry, and diffusion as an rf power transistor, so that it will thermally track the transistor. Tracking is

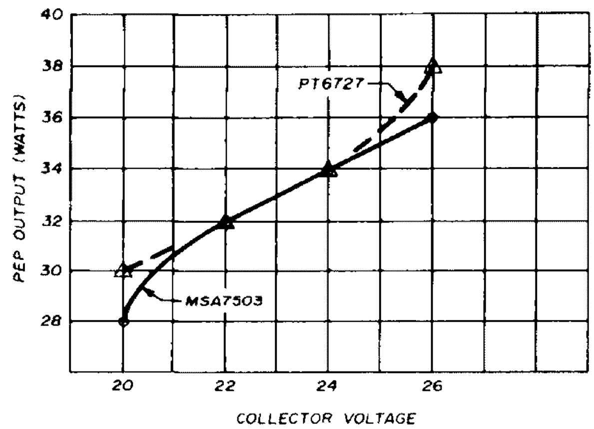


fig. 4. Outputs obtained from the Fairchild MSA7503 and TRW PT6727 transistors, plotted against collector supply voltage. These curves are not to be construed as indicating relative gains, since the drive power and tuning were optimized for each transistor at each value of collector voltage.

further improved by the temperature characteristics of the silicon resistor.

A constant current of approximately 350 mA is applied to injector terminal *I*, causing the diode to act as a voltage source having about 0.3 ohm impedance. The silicon resistor adds approximately 0.7 ohm and increases the apparent source impedance to approximately 1 ohm at supplier terminal *S*. The voltage at the supplier terminal will be between 0.45 and 0.85 volt, depending on the current being drawn from *S* and the temperature of the device. Thus, if a variable resistor is connected between the supplier (*S*) and reference (*R*) terminals, the supplier voltage can be adjusted. This is accomplished by the 4.7- and 100-ohm resistors shown in **fig. 5**; a single 5-ohm adjustable resistor could be used, but a 4.7-ohm, half-watt resistor in parallel with a printed-circuit type trimmer potentiometer provides finer control.

As the temperature of the byistor increases, the resistance of the silicon

table 2. Inductors and capacitors used in the amplifier circuits of figs. 3 and 5. Numbers in parentheses following the capacitance values are Arco/El Menco part numbers.

Q101	MSA7503	PT6727
C101	1.5-20 pF (402)	7-100 pF (423)
C102	7-100 pF (423)	24-200 pF (425)
C103	same as C102	3-35 pF (403)
C104	same as C102	2-25 pF (421)
L101	½ turn no. 18, 3/8" ID, 1¼" lead length	copper strap, 1" long, 3/8" wide
L102	1 turn no. 14, 3/8" ID, 1¼" lead length	3 turns no. 14, ¼" ID, ½" long
L103	7 turns no. 20, 3/16" ID, 3/8" long	7 turns no. 20, 3/16" ID, 3/8" long
L104	35 turns no. 20E wound on Micrometals T80-2 core	

resistor increases and the diode voltage decreases. This results in an increase in the apparent source impedance and lowers the bias voltage at the supplier terminal. Consequently, the base current of the associated transistor is reduced, preventing thermal runaway and providing improved dc stability of the amplifier. A more rigorous explanation of the byistor, with temperature-characteristic curves, appears in reference 5.

Aside from the biasing arrangement, the amplifier circuits of **figs. 3** and **5** are identical. Different values of inductance appear in **fig. 4**, plotted against collector supply voltage.

power supplies

The low-power stages require a 12- to 12.6-volt dc source which is capable of supplying approximately 1.5 amperes at peak power output. The MSA7503 final amplifier draws about 2 amperes, while the PT6727 requires a 2.5-ampere supply, both values being the peak current. Both of the supplies must be reasonably well regulated because of the varying load inherent in ssb operation.

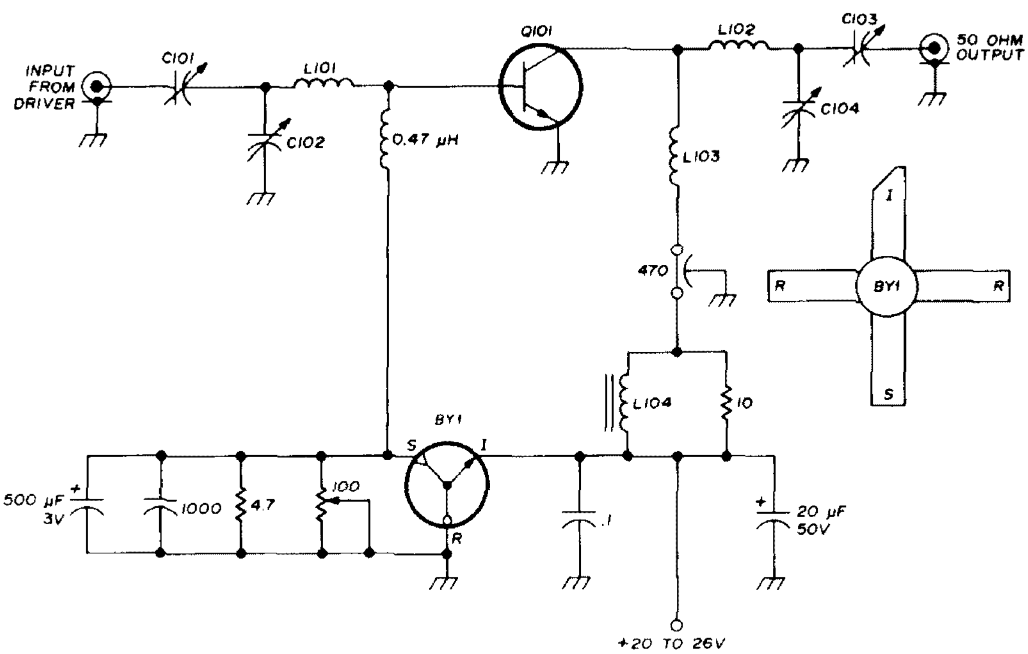


fig. 5. Schematic diagram of the final amplifier which has its base bias controlled by the CTC byistor. Details of parts identified by reference designators appear in table 2.

and capacitance in the input and output networks are required for each type of transistor, but either transistor can be used in either circuit. However, the improved construction of the PT6727 permits class-AB operation, which reduces the intermodulation distortion products to some extent. **Table 2** contains inductance and capacitance data applicable to either circuit, for each type of transistor.

The PT6727 appears to be somewhat better than the MSA7503 in terms of power gain, output, and distortion products, which is to be expected in view of its intended application. The power outputs obtained from the PT6727

A convenient way to obtain the two supply voltages is to use a 20- to 26-volt supply capable of providing the total load current, and incorporate a simple regulator circuit to drop the voltage to the nominal 12 volts required for the driver unit. Such a regulator is shown in **fig. 7.** The value of dropping resistor R will depend on the input supply voltage, and may be calculated from the equation shown on the diagram.

If the 20- to 26-volt supply is regulated with a circuit similar to or better than that shown in **fig. 7,** the 12-volt regulator is more than adequate for local-oscillator stability. Purists may want to add a 10- or 11-volt zener diode at the

local oscillator for additional regulation, but it was found to be unnecessary.

construction

Construction of the driver unit and the final amplifiers is shown in the various photographs. I started out using a piece of single-sided copper-clad board approximately 6-1/2 by 9-1/4 inches, since the circuits were developed stage by stage. I ultimately ran out of board, so for that reason the driver stage runs at a right angle to the low-level circuits. This is no problem except for the fact that it leaves a large portion of the board unused. To run all of the stages in the driver unit in a conventional straight line, I suggest using a piece of board approximately 12-inches long by 4-inches wide.

The normal techniques used for vhf construction should be followed—short leads and small, high-quality components. The low-level stages are each enclosed within shielded partitions which are made of pieces of copper-clad board soldered to the main board. The partitions should be placed across the transistor sockets to isolate the input and output circuits, thus minimizing any tendency of the high-gain stages to oscillate on their own. Liberal use of feedthrough capacitors and rf chokes for the supply voltages, with the dc wiring run on the top side of the board, prevents stray coupling through the power leads.

L1 and L2 in the mixer output circuit are shielded from one another by placing L1 and its associated capacitor on the mixer side of a shield partition, and L2 and its capacitor on the other side. The 0.47-pF coupling capacitor is then connected to the top end of each coil via a feedthrough terminal in the partition.

The MSA8507 (or B12-12) and PT6727 transistors are in stripline-opposed-emitter packages, which require some care in mounting. Virtually all of the published articles employ this package configuration in circuits which use printed-wiring inductances or transmission-line sections. Since my design uses only discrete components, the mounting and connection techniques are slightly different.

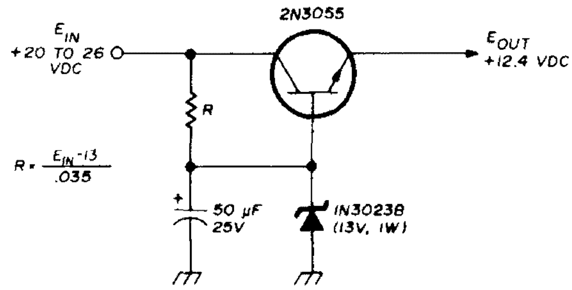


fig. 7. Regulator circuit for use in dropping the 20- to 26-volt dc supply to 12.4 volts.

There are two major conditions which must be met when mounting stripline transistors: the emitter leads must be grounded as closely as possible to the case, in order to minimize emitter lead inductance, and the case must be mounted on a heatsink without putting undue strain on any of the transistor leads. Considering the latter condition first, it can be satisfied by mounting the transistor to the heatsink, through a hole in the copper-clad board, *before* soldering to any of the leads. A sparse application of silicone thermal compound should be used between the body of the transistor and the heatsink.

Reducing the emitter lead inductance, as accomplished by soldering the leads close to the case, creates the annoying problem of what to do with the collector and base leads. Fortunately, operation at 144 MHz is not so critical as to preclude using one of the arrangements shown in **fig. 8**. In **fig. 8A**, the base and collector leads are soldered to lands which are insulated from the ground plane. These lands may be formed in one of two ways.

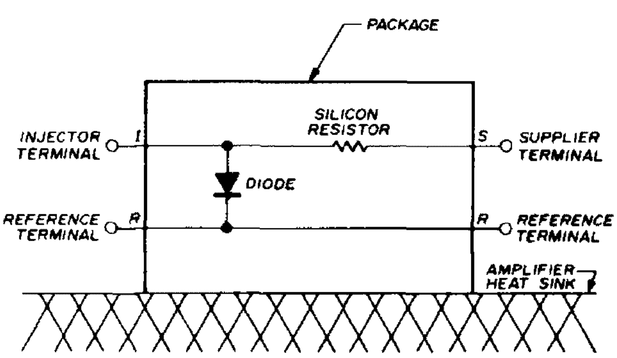


fig. 6. Schematic diagram of the byistor (courtesy CTC).

The copper can be routed out around the transistor leads, creating areas that are isolated from ground, and the leads then soldered to these lands. An alternate method is to cut small pieces of copper-clad board and cement them to the ground plane to form small insulated

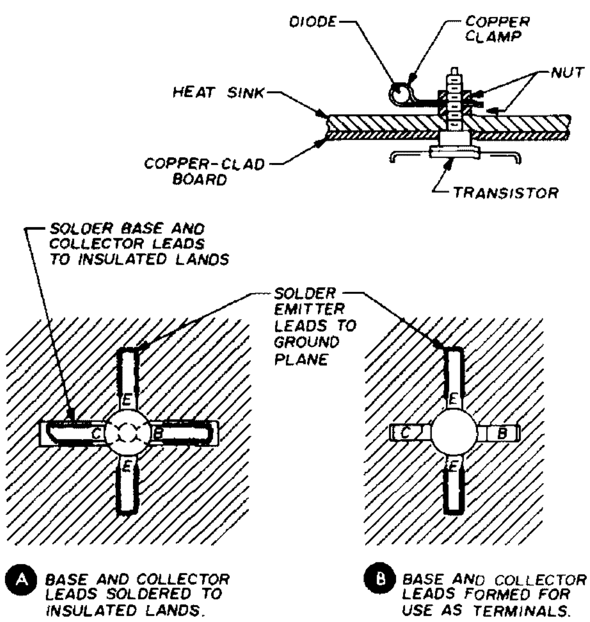


fig. 8. Methods of mounting stripline-opposed-emitter packaged transistors. Also shown is a base-bias control diode clamped to the transistor mounting stud for thermal tracking.

platforms to which the transistor base and collector leads can be soldered. In both cases, the heatsink must be spaced away from the main board so that the emitter leads are level with the ground plane or close enough to the ground plane so that they can be bent down slightly without too much strain.

Fig. 8B shows a third method which allows the heatsink to be mounted directly to the board without spacers. The base and collector leads are folded back on themselves, by means of long-nose pliers, and the folded ends carefully bent up away from the stud. This provides a relatively rigid terminal for connections to the transistor. The emitter leads are carefully bent down to the ground plane and soldered.

The heat sink for the driver transistor was made from a scrap piece of aluminum and has an area of about 6-1/2 square inches. This is enough radiating surface to keep the transistor from getting any more than barely warm to the touch. Of course, any one of the many commercial heat sinks having equivalent radiating surface could be used.

The 1N4001 diode is thermally coupled to the driver transistor by means of a clamp mounted on the transistor stud, as shown in **fig. 8B.** The clamp is made of a small piece of sheet copper which is formed around the diode to fit snugly. The diode and clamp surfaces should be coated with a thin film of thermal compound before being secured to the transistor stud. The diode cathode is soldered to the clamp, which is grounded via the heatsink, while the anode lead is connected to the bias-adjust resistor through a feedthrough capacitor.

The final amplifier is built on another piece of single-sided copper-clad board which measures 4 by 5 inches. The heat sink, which has a radiation surface of 33.4 square inches, is an Archer 276-1360, available at Radio Shack stores. The PT6727 stripline-packaged transistor is mounted to the board and heat sink in one of the ways previously described.

The MSA7503 is packaged in a TO-60 stud-mount case, which poses an additional problem in securing a low-impedance emitter-to-ground path. The emitter is connected internally to both the case and a terminal pin on the body, but using the pin is not practical because of the high internal lead inductance. The scheme shown in **fig. 9** was finally reached, and should be a useful method for mounting any similar transistor. First mount the heatsink to the board and, using a number-9 drill, drill a 0.196-inch hole through the heat sink and board for the transistor mounting stud. Then disassemble the heat sink from the board and enlarge the hole in the board to a diameter of 1/2 inch. Remount the heat sink on the board.

Obtain a small piece of copper *foil* (the kind used for electrostatic shields between power-transformer windings) and cut out a disc 1 to 1-1/4 inch in diameter. Carefully cut a hole in the

center of the disc just large enough to clear the transistor stud. Apply thermal compound to that part of the heatsink which is accessible through the enlarged hole in the board. Place the copper foil on the stud and mount the transistor to the heatsink. (Note that thermal compound is not used between the copper foil and the transistor, in order to maintain a good rf path between emitter and ground.) Slit the edges of the copper foil, now protruding from the hole in the board, so that the foil can be pressed flat against the copper board, and solder it down. This results in a continuous ground plane from the board to the transistor emitter.

If a 1N4719 diode is to be used to control the bias, mount it to the transistor stud in the same manner as described for mounting the 1N4001 on the driver transistor. If you use the BY1 byistor, mount it in one of the ways described for stripline packages, except that there is no need for concern about lead inductance. I located the byistor stud 1 inch from the transistor stud, on a line with the base lead. This places it under the input inductor, which hides it in the photograph of the PT6727 amplifier.

adjusting and tuning the driver

One of the advantages of having the final amplifier separate from the driver unit is being able to tune up the low-power stages independently of the final. And since two relatively high-power transistors are involved, having to worry about just one at a time makes the process much easier.

Before making any power connections, set the 100-ohm adjustable resistor in the driver bias network for maximum resistance. Then temporarily break the collector supply circuit in the driver stage and insert a 0-50 or 0-100 mA meter between L8 and the supply. Be sure the meter will indicate only the collector current and not the current drawn by the 1N4001 diode. Connect a good 50-ohm load and power meter to the driver output connector.

Connect the 12-volt supply and apply power. The meter should read zero or close to it. Adjust the 100-ohm resistor until the collector current is approximately 20 mA; this sets the operating bias on the driver transistor. Next check the operation of the 2N3563 oscillator, using an electronic voltmeter and rf probe at the output, or a sensitive detector coupled to the collector tank circuit. Tune the oscillator for maximum output. If the circuit fails to oscillate, it may be necessary to experiment with the value of the emitter bypass capacitor.

Turn off the power supply and replace the milliammeter with a 0-1 ammeter. Now connect the hf single-sideband transmitter, tuned to 29 MHz, to the input connector of the driver unit. Be sure that you have enough attenuation between the transmitter and converter to limit the power at the mixer input to 0.5 mW. Reapply power to the driver unit and slowly insert carrier at the transmitter while watching the driver-stage collector

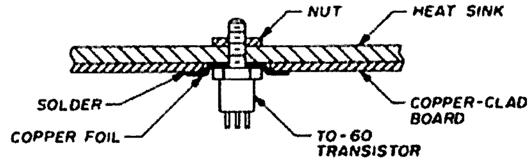
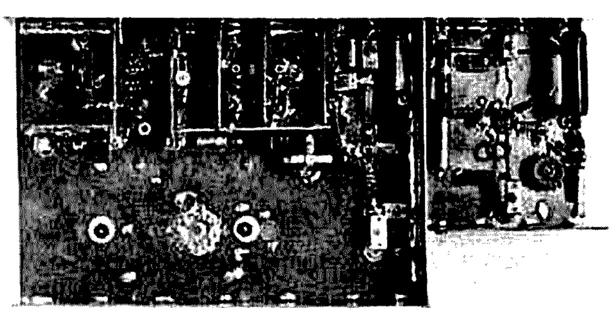
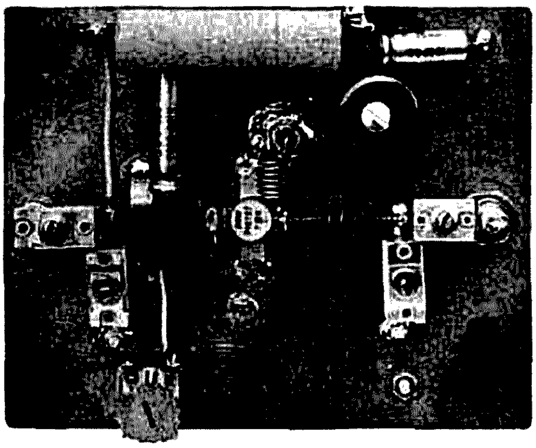


fig. 9. Method of mounting TO-60 stud-mount transistors to minimize emitter-to-ground lead inductance.



Bottom of the driver unit and MSA7503 amplifier. The local oscillator is at the left side, followed by the mixer and low-level stages to the right. The driver stage runs along the right side of the larger driver-unit board. The amplifier input circuit is at the top of the smaller board, and the collector circuit is at the bottom. Note the use of shield partitions to prevent feedback.

current. If the collector current starts to increase, immediately adjust the tuning capacitors in the driver output network for maximum output power. Actually, there is little likelihood of this occurring before the low-level stages have been tuned, so reduce the 29-MHz excitation



Component side of the PT6727 final amplifier. The copper-strap inductor in the input circuits hides the byistor. The transistor and byistor are mounted on the same type of heat sink as shown in the photograph of the MSA7503 amplifier.

and tune up the converter by means of the following technique.

Tune each stage for maximum power output. An electronic voltmeter with an rf probe, connected at a point which follows the circuit or stage being adjusted, makes a good tuning indicator without loading down the circuit (e.g., connect the probe to the collector of the stage *following* the one being tuned). As each stage is tuned, gradually increase the 29-MHz carrier and monitor the driver collector current so that the driver output circuit can be tuned for maximum output as soon as the collector current starts to increase. As excitation to the driver stage is increased, the collector current will rise to a maximum of 0.75 to 1 ampere. Tune the driver output circuit for maximum output consistent with minimum collector current. Since the Q of the output circuit is low, tuning is relatively broad, making it easy to pick the point of best efficiency.

As the 29-MHz drive is increased and as each stage is tuned, the output should gradually rise to at least 6 watts. However, if the output goes to 9 watts or so, it is an indication that one or more of the low-level stages are saturated. If this happens, reduce the excitation to the point where the output power drops sharply. This is the limit of linear operation, and all tuning adjustments should be repeaked at this level. Vary the frequency of the ssb transmitter from 28 to 30 MHz and retune it for constant output at several points within the frequency range, but do not retune the transmitting converter. The output from the converter should vary less than 10 percent.

Deenergize the power supply, remove the ammeter, and restore the driver collector circuit to its original state. You now have a 6-watt ssb signal, ready to put on the air or to drive the final stage. If you want to get it on two meters at this point, be sure to read the section headed *operation* before connecting the antenna.

adjusting and tuning the final amplifier

If you are using the amplifier circuit shown in fig. 3, set the 100-ohm adjustable resistor for *maximum* resistance. If you are using the circuit of fig. 5, set the 100-ohm pot for *minimum* resistance between the byistor supplier terminal and ground. Temporarily open the collector circuit, as was done for the driver, and insert a milliammeter between L104 and the power supply so that it will measure only the collector current. A 0-50 or 0-100 mA meter can be used for the PT6727; a 0-10 mA meter is preferable if an MSA7503 or equivalent is used.

Using the lowest supply voltage which will provide you with the output power that you need, turn on the power supply and adjust the bias resistor on the amplifier for a collector current of 25 mA if the PT6727 is being used. If you are using an MSA7503 or an equivalent transistor, adjust the bias resistor to the point where the collector just starts to draw current—about 2 or 3 mA. Remove power and replace the milliammeter with an am-

meter having at least a 2.5 ampere range.

Connect the driver unit to the amplifier by means of a short length of 50-ohm coax cable, and terminate the amplifier with a power meter and good 50-ohm load. Energize the driver unit and amplifier power supplies, and *gradually* apply rf excitation. Tune the amplifier input and output capacitors for maximum output each time the drive is increased. The output should rise smoothly until it reaches the approximate value indicated in **fig. 4** for the supply voltage being used. As with the driver stage, the final tuning should provide maximum efficiency (maximum output consistent with minimum collector current). The collector efficiency of the PT6727, operating class AB, should be approximately 60 percent. The efficiency of the MSA7503 or any other transistor operating virtually at cut-off may be as high as 75 percent. Tuning the exciter over a 2-MHz range should not affect the output of the transmitting converter by more than 10 percent.

Driver-stage tuning may be refined during the amplifier tuning procedure by peaking the capacitors in the driver collector circuit for maximum amplifier output, but this must not be done until after the amplifier input tuning capacitors have been adjusted for maximum output. Then remove all power, disconnect the ammeter, and restore the final collector circuit to its original condition.

operation

The transmitting converter is now ready to feed an antenna or to drive a high-power linear amplifier. In the latter case, connect the converter to the amplifier through a 50-ohm coax cable (assuming that the amplifier being driven has a 50-ohm input impedance) and retune the converter amplifier collector circuit for maximum drive. It is advantageous to monitor the transistor amplifier collector current to achieve maximum efficiency, which can be done simply by inserting an ammeter in the lead from the dc supply. Remember, however, that you will now be measuring the collector current plus the current drawn by the bias-control diode or byistor, so that the total current through the meter will be 200 to 350 mA greater than the collector current alone.

If the transmitting converter is fed directly to an antenna, a lowpass filter must be inserted between the output

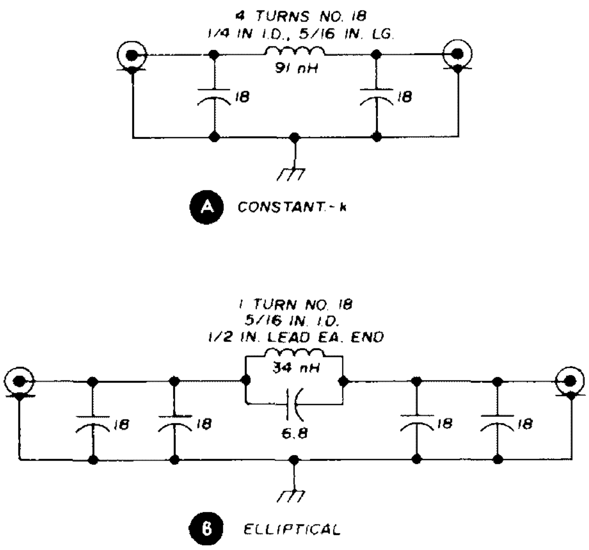


fig. 10. Schematic diagrams of two lowpass filters for suppressing harmonic radiation. The coil in the elliptical filter (B) should be adjusted so that it resonates with the 6.8-pF capacitor at 327 MHz. Capacitors are silver mica, 5 percent or better.

connector and the transmission line to attenuate harmonics which will be passed by the low-Q output network of the driver or amplifier. (The higher-Q tuned circuits in a vacuum-tube amplifier following the converter will provide sufficient filtering, and eliminate the need for a lowpass filter.) Two such filters are shown in **fig. 10.** The constant-k pi-section in **fig. 10A** is slightly simpler than the elliptical pi-section of **fig. 10B,** but the latter will provide at least 6-dB, and as much as 16-dB, more attenuation to the second harmonic than will the constant-k configuration.

After making the necessary connections and applying power, retune the output collector circuit for maximum output power. Again, it is wise to monitor the collector current, as described above. Once the preceding tuning procedures have been completed, it will not be necessary to retune any of the circuits

appendix

Most of the parts used in the transmitting converter are available through regular distributors. The following list is provided for those items which must be ordered from other sources, and includes prices (as of July 1973) for those of major importance.

item	unit price	source
CTC B12-12 BY1	\$ 9.50 6.00	Communications Transistor Corporation, 301 Industrial Way, San Carlos, California 94070
Anzac MD108	7.00	Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154
TRW PT6727	35.00	Request name and address of closest distributor from Marketing Department, TRW Semiconductor Division, 14520 Aviation Boulevard, Lawndale, California 90260
Fairchild FMT4170	5.50	Request name and address of closest distributor from Marketing Department, Fairchild MOD, 4001 Miranda Avenue, Palo Alto, California 94304
Micrometals cores		Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607

over long periods of time, provided that you do not change the load or supply voltages. The low-Q tuned circuits are relatively insensitive to other changes.

conclusions

Operation on two meters during the past several months, using both amplifiers, has shown that the transmitting converter is stable and trouble-free. A spectrum analyzer was not available for distortion measurements, but rough measurements using a receiver and calibrated step attenuator indicate that the third-order products are down approximately 24 dB when using the MSA7503 amplifier, and approximately 27 dB for the PT6727. The limitation in the latter case is probably due to the distortion products generated in the MSA8507 driver stage.

acknowledgements

I would like to acknowledge the technical and material assistance received from the following: Joe Reisert, W6FZJ, of Fairchild Microwave and Optoelectronics Division; Jack Manon, W6FIG, of TRW Semiconductors; and Bob Artigo, W6GFS, Lee Max, and Mike Mallinger of Communications transistor Corporation. Thanks are also due to Alan Stein for the photography.

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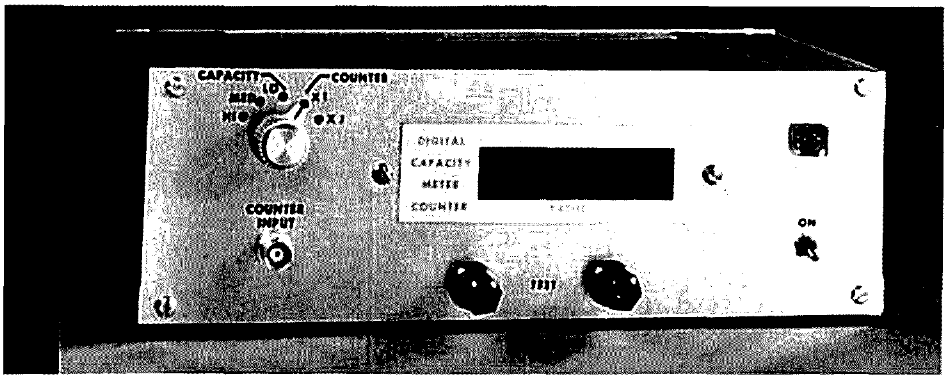


digital capacitance meter

Construction details for a wide-range digital capacitance meter that doubles as a 20-MHz frequency counter

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

Depending on your point of view, this instrument may be called a capacitance meter which will also function as a frequency counter or it can be called a frequency counter which will also measure capacitance. To me it's a capacitance meter since that was my need at the time I designed it. However, to provide one function without the other would be foolish since circuitry for both is practically identical and requires only the switching of a few points in the control logic to implement either mode of operation.

theory of operation

The capacitor to be tested is placed in a timing circuit whose output gates a train of fixed-frequency pulses into a standard counter. The output pulse length from the timer circuit is proportional to the size of the capacitor, thus varying the gate time. The resultant count is indicated by the digital readouts. A large capacitor would result in a long gate time and a high pulse count.

If the resistance factor in the RC time constant is used as a calibrating device, it could be adjusted in conjunction with a known value of capacitance to give a known gate time and, therefore, a known count. For example, if R were adjusted to provide a 1.0-millisecond output pulse in conjunction with a 1000-pF capacitor and the pulse rate was 1.0 MHz, during the 1.0-ms opening of the gate 1000 pulses would get through to the counter and register on the readouts. A 900-pF capacitor would shorten the gate time sufficiently to allow only 900 pulses through. Larger capacitors permit proportionately longer count times with resulting higher counts.

In the capacitance meter frequency is fixed and gate time is variable, while in the counter gate time is fixed and frequency is variable.

circuit details

About the time I first started thinking about this idea, Signetics introduced their NE555 IC timer. This little item requires only two external components, a resistor and a capacitor, and is just the thing for

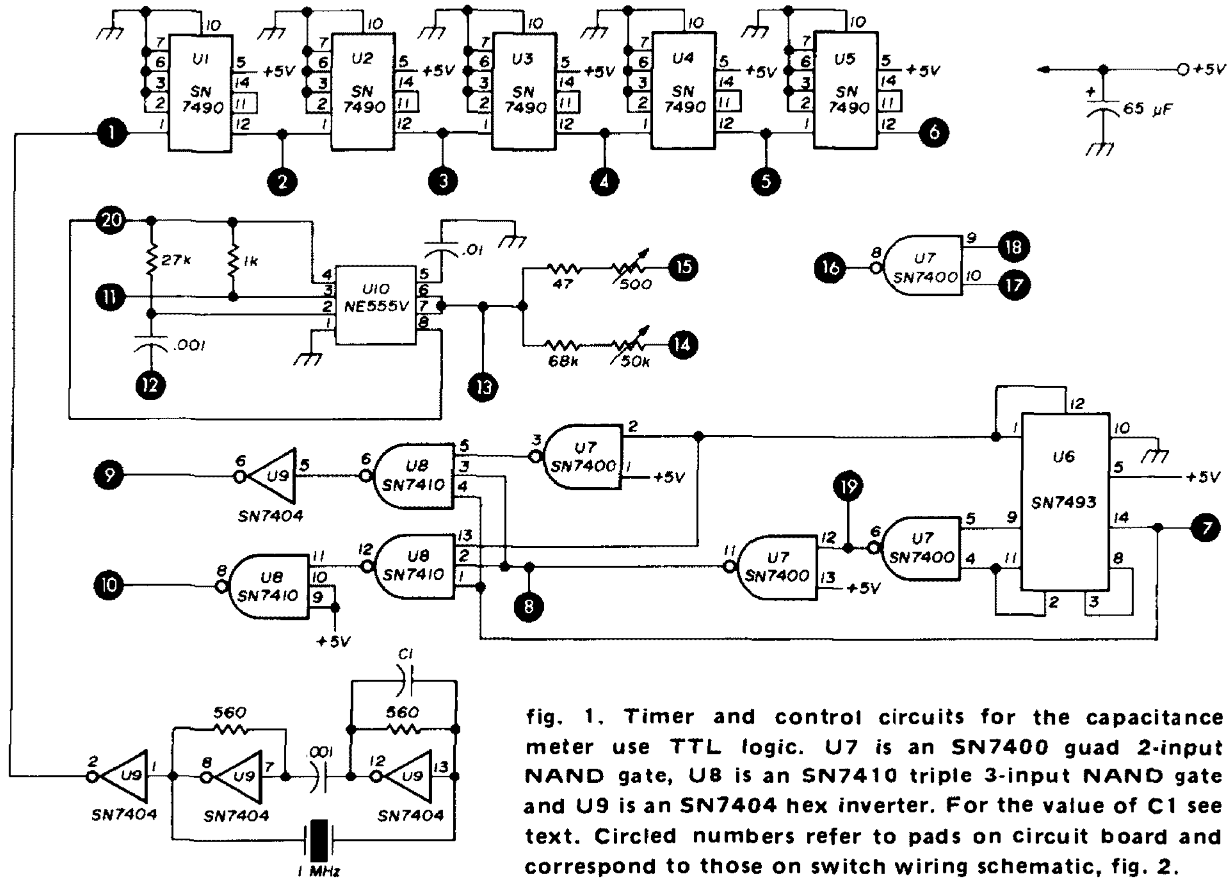


fig. 1. Timer and control circuits for the capacitance meter use TTL logic. U7 is an SN7400 guad 2-input NAND gate, U8 is an SN7410 triple 3-input NAND gate and U9 is an SN7404 hex inverter. For the value of C1 see text. Circled numbers refer to pads on circuit board and correspond to those on switch wiring schematic, fig. 2.

generating the timing pulses. In this case the resistor would be a calibrating pot and the capacitor would be the unit under test.

Inside the NE555 are two comparators, a flip-flop, an output stage and a discharge transistor. Initially, the capacitor is held discharged by the transistor connected across it. When a negative-going pulse is applied to pin 2 of the IC, the flip-flop is set, releasing the short across the capacitor and charging commences.

A circuit operating in this mode is the old familiar one-shot or monostable. The NE555 may also be wired as an astable if free-running operation is desired. For interested readers, the data sheets show many other interesting applications for this IC.

The control logic circuit used in this instrument was borrowed from an article by W1EO in *QST.*[1] A 1-MHz crystal oscillator and SN7404 hex inverter IC were added to provide the clock input (see **fig. 1**). Five SN7490 decade counters

table 1. Capacitance ranges used in the instrument built by the author.

range	calibration	clock frequency	readout format
1000 µF	1.0 µF = 0.1 ms	100 kHz	1000.0
1.0 µF	.001 µF = 0.1 ms	100 kHz	1.0000
0.1 µF	.001 µF = 0.1 ms	1.0 MHz	0.1000

The reference voltage for the comparator is internally set at two-thirds of the operating voltage. When the voltage ramp across the capacitor reaches this level, the circuit fires, resetting the flip-flop and discharging the capacitor. Upon receipt of another trigger pulse, the cycle repeats.

divide the crystal frequency down to 10 Hz — this results in a string of pulses spaced exactly one-tenth second apart. An SN7493 is used as a divide-by-twelve counter to provide a period of 1.2 seconds or 12 clock pulses for a complete timing cycle.

The initial 1-second portion is the count period during which the count gate is open. During the 0.2-second interval between counting periods, a transfer pulse is generated which allows the information stored in the latches (if used) to be shifted to the decoder/drivers for readout of the latest count.

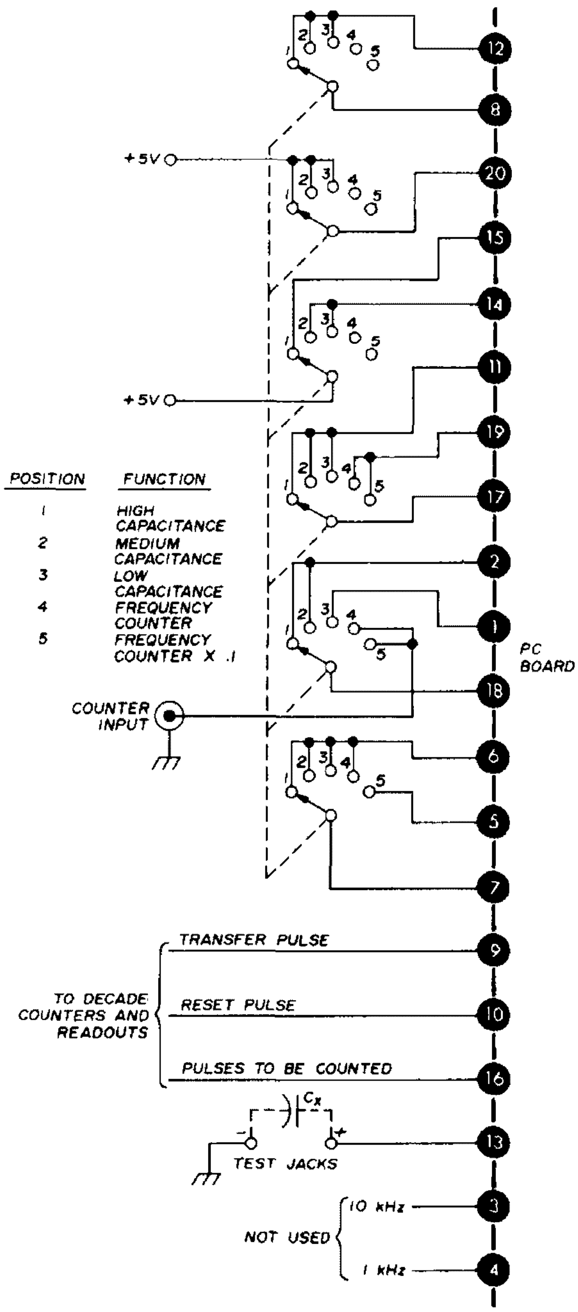


fig. 2. Wiring diagram for the function switch. Circled numbers refer to pads on the circuit board (see fig. 4).

A subsequent reset pulse is also generated during this interval which returns all counters to zero in preparation for the next 1-second counting period. These latter two pulses are formed by interconnecting various gates contained in an SN7400 and an SN7410 IC. The pulse appearing at pin 11 of U7 is negative-going at the start of the timing period and is used to trigger the NE555 for capacitance measuring.

When the instrument is operating as a capacitance meter, the control pulse for the count gate comes from the timer circuit and the pulse train to be counted is generated by the internal clock. When functioning as a straightforward counter, the count gate reverts to internal control and the signal to be counted comes from an external source. These and other points require switching and are combined into a single multi-pole switch. In my unit this switch provides three capacitance ranges and two for counter operation. Sections of the function switch are also used to apply power and trigger pulses to the timer when operating in the capacitance-measuring mode (see fig. 2).

Table 1 shows the relationships between the various parameters when applied to a 5-digit counter such as that used here. Obviously this scheme is not a mandatory one and can be altered to suit other situations. If you are planning to place decimal points at appropriate points in the display, don't forget to reserve a pole on the function switch for that purpose.

construction

The heart of the capacitance meter is the control logic and timer circuitry. A two-sided PC board was laid out to accommodate all of the circuitry in an uncrowded area 2.5 by 4.6 inches. Since the TTL logic ICs come in dual-inline packages, a similar version of the Signetics NE555 timer was used. This is their 8-pin mini-DIP known as the V package (NE555V).

The 1-MHz crystal is in an HC6-U holder with wire leads. The calibrating trimmers are the common 1-inch type which have pin spacings of 0.2 and 0.3 inch with a 0.2-inch stradle. The decoupling filter capacitor is a 65-μF dipped

tantalum but any substitute unit of 50-μF or so may be used if it fits on the board.

Circuit pads are provided at all points being switched as well as at inputs and outputs. A pad is provided at the crystal output as well as at each decade although not all frequencies will be used in this particular application.

The most practical scheme I've been able to devise for making double-sided PC boards is to cut out a cardboard frame for the blank board so that it just fits in the opening without moving around. Position one of the negatives over the opening with the proper side up and tape the edges to the cardboard frame. Turn the frame over and place the second negative so that the two are back-to-back and in perfect registration. Tape this one along one edge so that it may be lifted to allow insertion of the blank PC board. After both sides have been exposed, develop

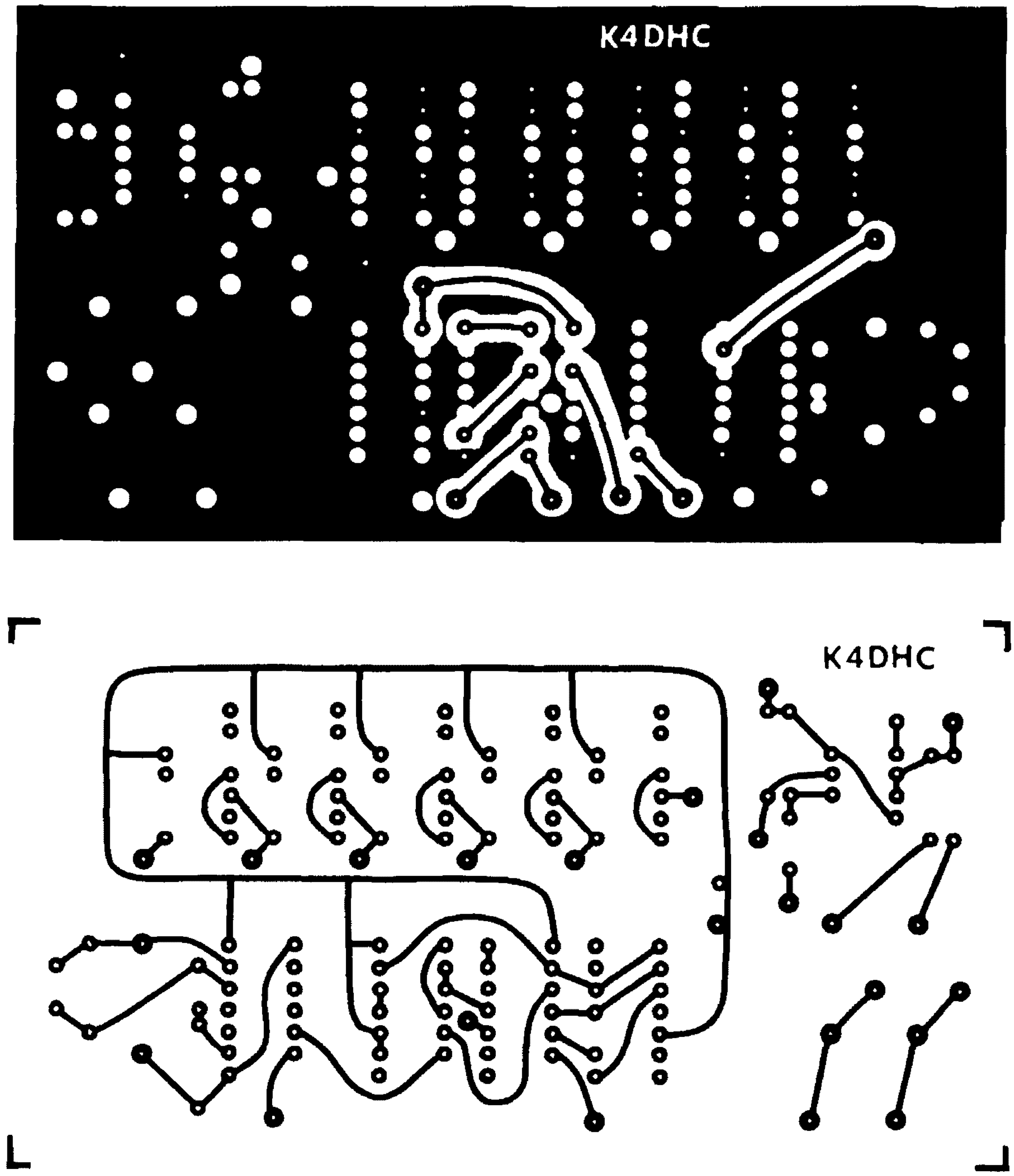
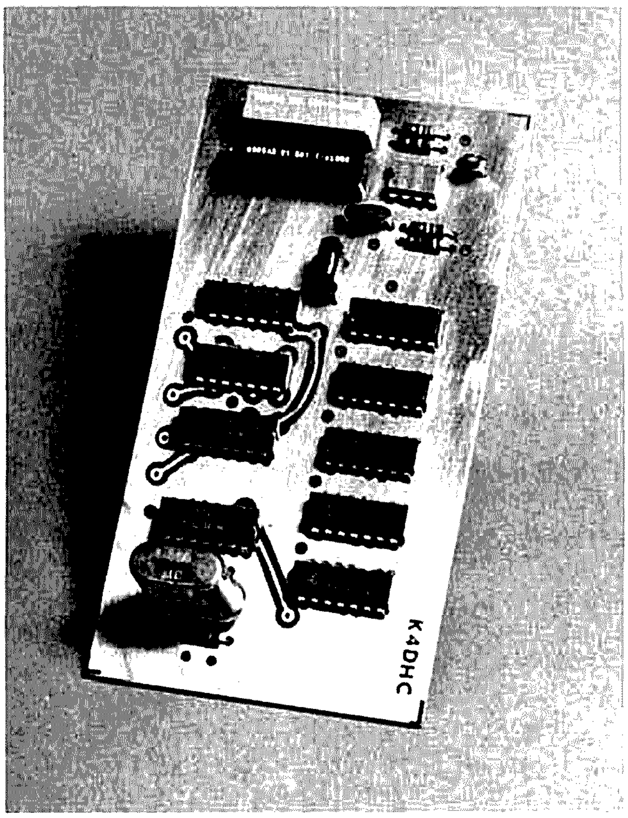


fig. 3. Full-size layout for both sides of the double-sided circuit board.

and etch in the normal manner. The board in the photograph was homemade in this way (see **fig. 3** for the layout).

If you are planning to build one of these instruments from scratch, you'll need several decades of counting and readout circuitry. Many of the advertisers in *ham radio* sell kits consisting of a counter, a latch, a decoder/driver and a readout device, along with a PC board for easy assembly. Four decades would be the minimum required. Anything over that would be at the builder's discretion.



Component side of the printed-circuit board containing the timer and control circuits for the digital capacitance meter.

I used five stages because I happened to have five hybrid assemblies on hand which were suitable for this application. Each of these dual-inline packages contained a counter, a latch and a decoder/driver. I mated these with five homemade readouts and ended up with a neat 5-digit counter section.

I would have used one or two more stages if I'd had more of the hybrids since it would have made the frequency counter a little more useful. For capacitance measurements, however, the five digits are adequate since the accuracy of the system doesn't really warrant any greater resolution.

If you already own a counter and don't mind tearing into it, you could do a little rearranging along the lines described here to add the capacitance measurement feature. In counters that provide for external gate control the output from the timer could be fed into this connection. In addition, a suitable trigger pulse must be brought out to fire the timer at the start of the cycle. Suitable clock pulses could also be brought out for the various operating ranges.

It is by no means mandatory that a PC board be used for assembly. The circuit described here was at one time made up on a piece of perforated board and wired from point-to-point. It worked just fine.

calibration

All you need for calibration are a couple of fairly close tolerance capacitors of suitable values. With a capacitor connected to the test jacks and the instrument switched to the high range, adjust the 500-ohm trimmer for proper display of the value. Adjust the 50k trimmer for either of the two remaining ranges.

A capacitor of around 1.0 μF could be used for setting both trimmers since there is an overlap between ranges. The more points you can check, of course, the more accurate the instrument. From my experience it seems reasonable to expect at least 10% accuracy across the operating range of 1000 pF to 1000 μF.

Since this unit was intended primarily to measure large capacitors, readings should be close enough for most experimental work. They will also bear out the fact that most electrolytics have values a lot higher than marked.

It should be pointed out that the unit will read well over 1000 μF but accuracy falls off rapidly above 1500 μF. This is apparently due to shortening of the output pulse from the timer as the duty cycle increases. At the opposite end, reading values below 1000 pF seems to be

impractical due to bad jitter on the timer output pulse. The comparator input which the capacitor is connected across is a high impedance point and consequently picks up *all* kinds of *noise* and hum. Looking at the trailing edge of the output pulse on a scope will verify this. The end result is that the count gate sees a constantly varying count time which makes valid readings impossible.

All circuitry is powered from a single 5-volt supply capable of supplying the required current. In my instrument maximum current is about 1.5 amperes. Close regulation is not essential as voltage variations will not affect the timer output. When you are measuring electrolytic capacitors, remember that they should have a minimum rating of 6 volts just to be safe.

summary

Parts of this circuit may be of interest to some readers even if not all of it is. The control logic may be suitable for a counter you've been thinking of building or the timer circuitry may be extracted for use with an existing counter. The crystal oscillator could be modified for 10-MHz operation by adding another decade divider. A calibrating trimmer capacitor could also be added in series with the crystal for precise adjustment of the clock. This would be primarily for improving frequency measuring accuracy..

Incidentally, you may find that some 1-MHz crystals won't oscillate at their fundamental frequency. A scope should be used to check this. Holes have been provided on the circuit board to install a capacitor across one of the feedback resistors if this problem is encountered. Try about 100 pF as a starting value and substitute values until you're sure the oscillator will start properly every time you fire up.

A preamplifier and conditioning circuit for the counter was not included on the board. There have been numerous examples of such circuits in all the amateur publications so finding what you want should not be too difficult.

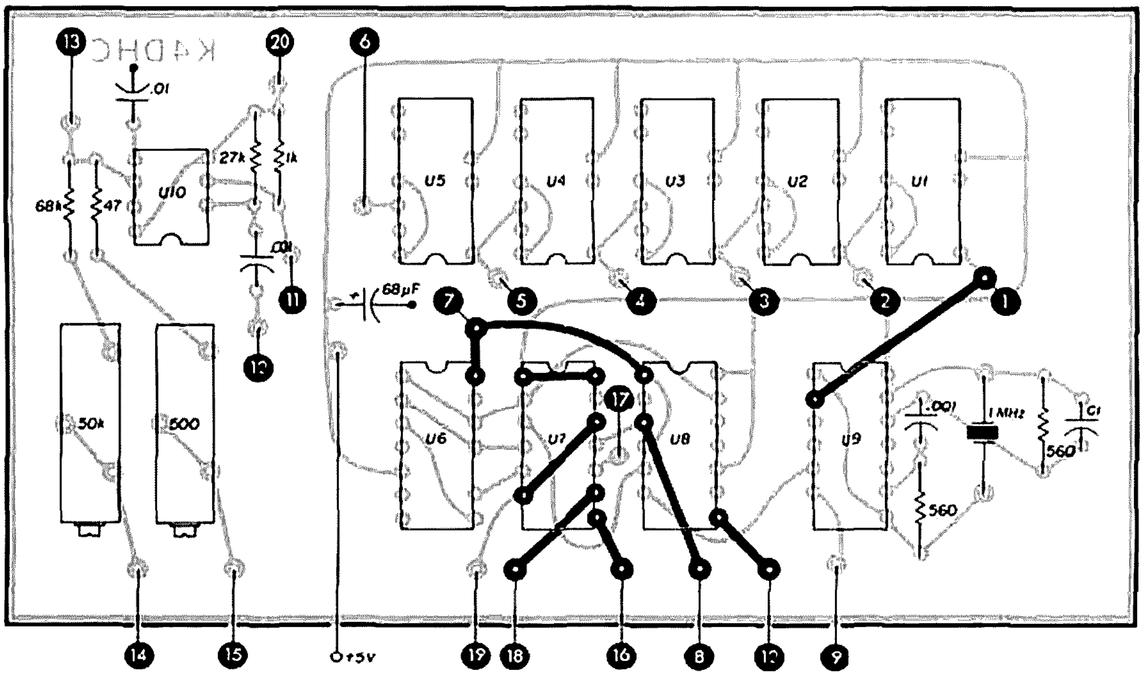


fig. 4. Component layout for the capacitance meter circuit board. Circled numbers correspond to connections on the function switch (see fig. 2). Bold traces are on component side of the printed-circuit board.

reference

1. Kenneth Macleish, W1EO, "A Frequency Counter for the Amateur Station," *QST*, October, 1970, page 15.

ham radio

how to design L-networks

How to choose the proper L-network for your particular impedance-matching problem, and how to calculate the required component values

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

Graphical methods of designing L-networks have been presented several times in the past.[1,2,3] As shown in reference 1, there are eight possible L-networks for matching a pure resistance to *any* impedance. In most amateur cases the pure resistance is the 52-ohm coaxial transmission line, and the impedance is that at the base of a vertical antenna.

It is important to note that only certain networks can be used to match certain ranges of impedance. Also, because of possible mutual coupling, networks using two inductors are less desireable than the others. The lowpass filter network is the most desireable, but can be used only for some load conditions. One criteria which affects the choice of network is whether or not the antenna resistance is greater or less than 52 ohms. A more definite way of selecting the correct network is shown in the graphs that follow.

One of my former graduate students, John Lewis, studied the L-network situation and found that there are three different networks that will match any conceivable load impedance. He developed equations for these three networks and wrote a computer program that would, for a given problem, select the proper network and calculate the two necessary L-network element values. This article will give those equations, and describe them so that you can design your own L-networks, using simple equations and elementary arithmetic.

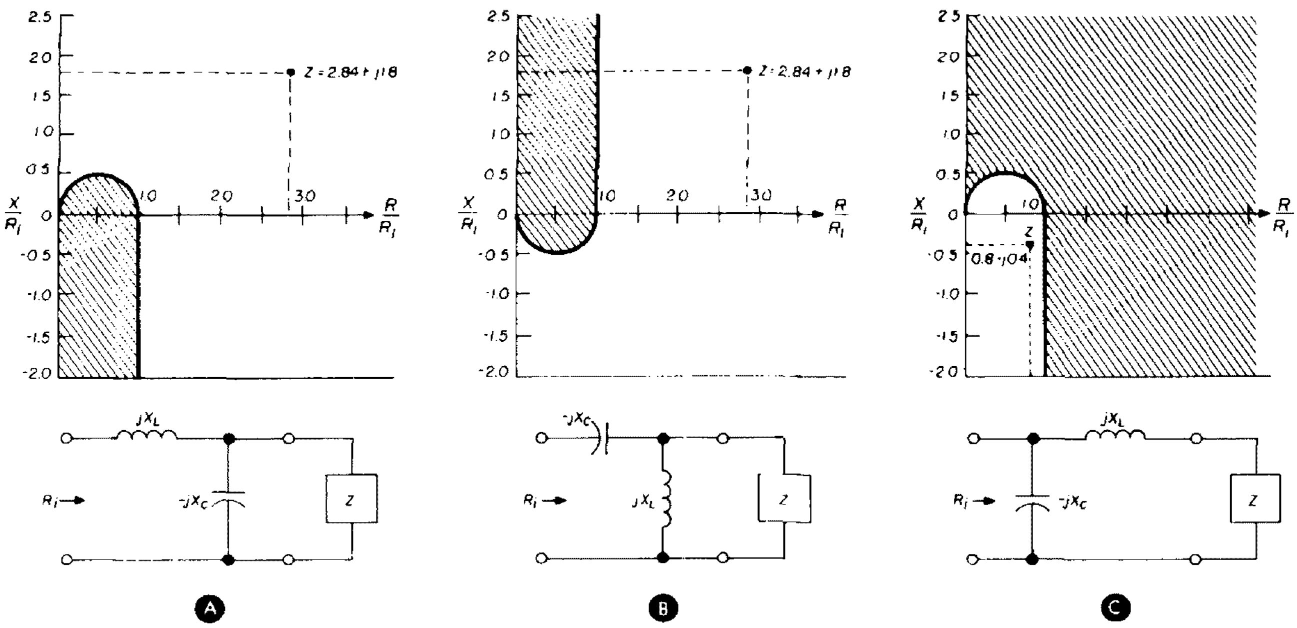
He developed the equations by writing the network equations for a given network, calling the input impedance R_i (50

ohms in our case). He then solved the equations for the network element values. For example, for the case of the network in **fig. 1C**

$$R_i = \frac{-jX_c\ (R + j(X_L + X)}{R + j\ (X + X_L - X_C)}$$

tions for X_L and X_C. John called these networks **A, B,** and **C** as shown in **fig. 1.**

The best way to show which load impedances each network can match is by means of a graph, first presented by Smith in *Electronics.*[4] The shaded part of each graph shows those load values which that



Network A $$X_C = \frac{-2\,R_i\,X + k}{2(R - R_i)} \qquad (1)$$

$$X_L = \frac{R_i\,X + X_C\ (R - R_i)}{R} \qquad (2)$$

Network B $$X_L = \frac{2\,R_i\,X + k}{2(R - R_i)} \qquad (3)$$

$$X_C = \frac{X_L\ (R - R_i) - R_i\,X}{R} \qquad (4)$$

Network C $$X_L = \left(\sqrt{R_i\,R - R^2}\right) - X \qquad (5)$$

$$X_C = \frac{(X_L + X)^2 + R^2}{X_L + X} \qquad (6)$$

$$\text{constant } k = \sqrt{4\,R_i^2\,X^2 + 4\,R_i\,(R - R_i)\,(X^2 + R^2)} \qquad (7)$$

fig. 1. Three types of L-networks which may be used for impedance matching. The accompanying graphs show the range of impedances which may be matched by each of the networks. Point Z in (A) and (B) is the normalized impedance used in the first example in the text. Point Z in (C) is the normalized load impedance used in the second example.

solving, $X_L = (R_i\,R - R^2)^{½} - X$

$$X_C = \frac{(X_L + X)^2 + R^2}{X_L + X}$$

The steps in the solution are not shown here. To do that the first equation was expanded and the real and imaginary terms properly equated, resulting in solu-

particular network cannot match. The network can provide a match for any impedance in the non-shaded area. The graphs are normalized, which means that all graph values are divided by the impedance value of transmission line used (50 ohms). Thus, a load resistance of 50 ohms shows up on the graph as 1 unit horizontally.

When using the graphs and formulae presented in **fig. 1,** solve first for the net-

work element given in the left-hand column. For example, assume you have a vertical antenna with an input impedance of 142 + j90 ohms and want to feed it with 50-ohm coaxial cable. Therefore, R = 142 ohms, X = 90 ohms and R_i = 50 ohms. Normalizing, $R/R_i = 2.84$ and $X/R_i = 1.8$. In this case either network A or B must be used because the normalized impedance (Z = 2.84 + j1.8) falls into the forbidden region in the graph for network C.

To use network A, first calculate the constant, k, from **eq. 7**. Then find X_C and X_L, respectively, using **eqs. 1** and **2**.

$$k = \sqrt{(4 \cdot 50^2 \cdot 90^2) + (4 \cdot 50)\,(142 - 50)\,(90^2 + 142^2)} = 24516.48$$

$$X_C = \frac{-(2 \cdot 50 \cdot 90) + (24516.48)}{2(142 - 50)}$$

$$= 84.33 \text{ ohms}$$

$$X_L = \frac{(50 \cdot 90) + 84.33(142 - 50)}{142}$$

$$= 86.33 \text{ ohms}$$

To determine the component values for network B calculate X_L and X_C from **eqs. 3** and **4**, respectively. The constant, k, is the same as before.

$$X_L = \frac{(2 \cdot 50 \cdot 90) + 24516.48}{2(142 - 50)} = 182.15 \text{ ohms}$$

$$X_C = \frac{182.15\,(142 - 50) - (50 \cdot 90)}{142}$$

$$= 86.32 \text{ ohms}$$

These values check with the graphical solutions shown in **figs. 2** and **3** (see reference 3 for application with a 7-MHz vertical antenna).

As another example, assume that you want to match a 50-ohm transmission line to an antenna with an input impedance of 40 - j20 ohms. Therefore, R = 40 ohms, X = -20 ohms and R_i = 50 ohms; $R/R_i = 0.8$ and $X/R_i = -0.4$. The normalized input impedance is 0.8 - j0.4 ohms. This value can be matched by network C but

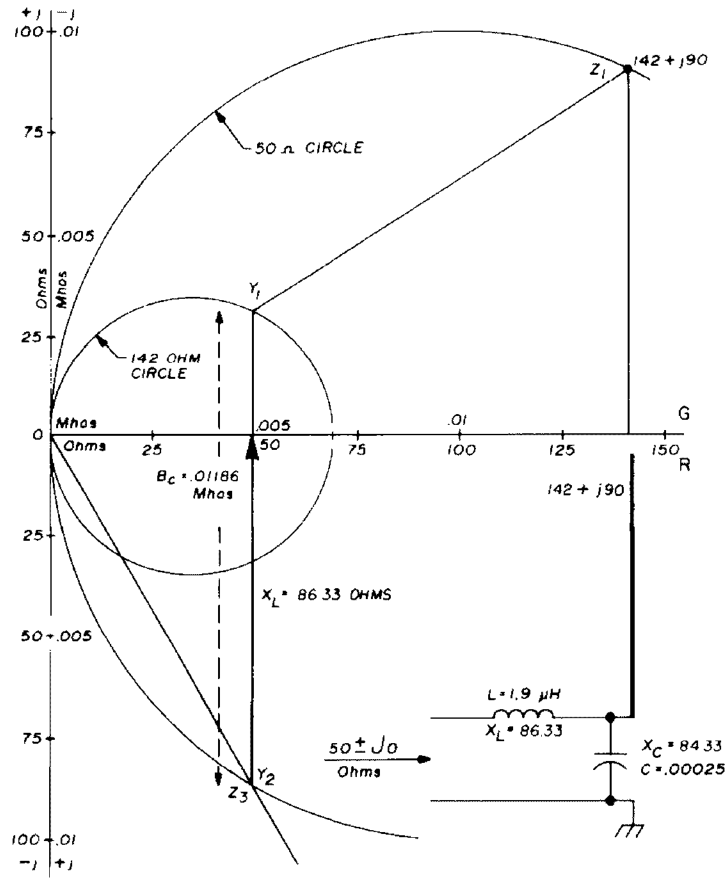


fig. 2. Graphical solution using the L-network of fig. 1A to match a load impedance of 142 + j90 ohms.

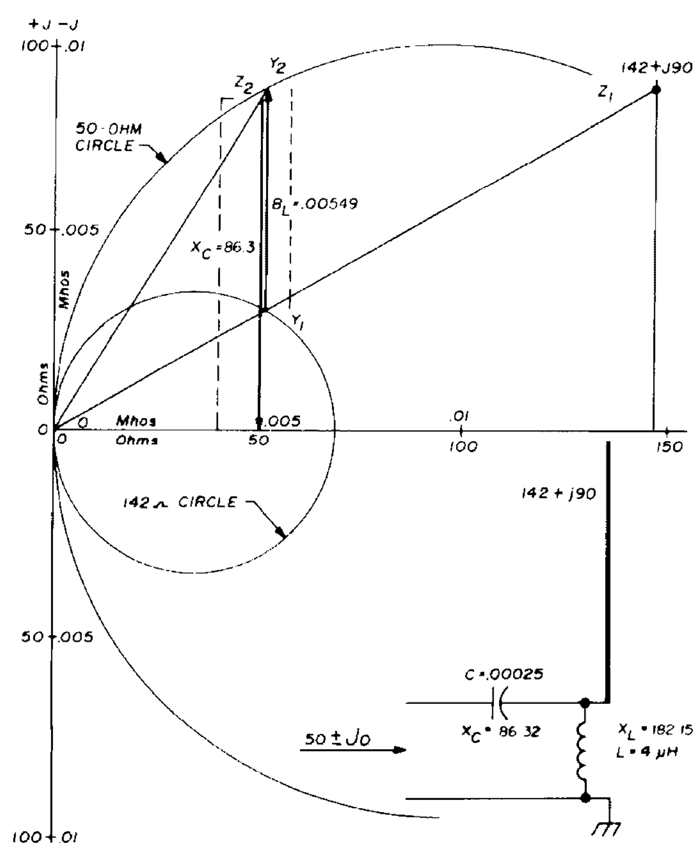


fig. 3. Graphical solution of the L-network of fig. 1B to match a load impedance of 142 + j90 ohms.

falls into the forbidden region in networks A and B.

To determine the proper values for network C first calculate X_L, using **eq. 5.** Then find X_C using **eq. 6.**

$$X_L = [(50 \cdot 40) - 1600]^{1/2} + 20 = 40 \text{ ohms}$$

$$X_C = \frac{(40 - 20)^2 + 1600}{40 - 20} = 100 \text{ ohms}$$

To check the correctness of these values it is necessary to calculate the impedance seen at the input terminals. From inspection, it can be seen that Z_C is in parallel with the series combination of Z_L and the complex load impedance Z. Using the formula for parallel impedances:

$$R_i = \frac{(Z_C)\,(Z + Z_L)}{Z_C + (Z + Z_L)}$$

$$= \frac{(-j100)\,(40 - j20 + j40)}{(-j100)\,(40 - j20 + j40)}$$

$$= \frac{(-j100)\,(40 + j20)}{(-j100) + (40 + j20)}$$

$$= \frac{-j4000 + 2000}{40 - j80}$$

Multiplying by the conjugate:

$$\left(\frac{-j4000 + 2000}{40 - j80}\right)\left(\frac{40 + j80}{40 + j80}\right)$$

$$= \frac{400 \times 10^3}{8 \times 10^3} = 50 + j0$$

This network provides a perfect match to 50-ohm transmission line.

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4. Phillip H. Smith, "L-Type Impedance Transforming Circuits," *Electronics,* March, 1942, page 48.

ham radio

RTTY message generator

Complete construction details for an RTTY message generator that uses TTL digital logic

C. A. Ellsworth, W6OXP, W. G. Malloch, W8KCQ

For a number of years, the users of teletypewriter services have relied on an automatic response from an interrogated RTTY terminal unit to confirm completion of a desired traffic circuit. The interrogated terminal, upon command, generates a station identification code or message such as, DE KX6IT. This message is usually generated by an electromechanical device consisting of a number of coded bars on a rotating drum, momentarily closing electrical contacts.

With the advent of low-cost, multifunction integrated circuits, it is feasible to generate the message using digital logic. This increases reliability and makes maintenance easier as well as lowering the cost. Moreover, some electro-mechanical message generators are mechanically peculiar to a specific type or family of teleprinters. The digital logic method is directly applicable to any machine or circuit of any family of teleprinters using compatible signaling codes.

RTTY signaling code

The presently used Baudot (Murray) RTTY code is a binary code, a two-state condition, such as the presence or absence of current. As applied to most teletypewriter circuits, it is a condition of current flowing in a loop (*mark*) or no current flowing in the loop (*space*). Each printed character or machine function is determined by the sequence of mark and space pulses received by the machine.

The format of the signaling code depends on the maximum number of different characters to be printed or functions to be performed by the machine. The two most common arrangements used are the 5-level and 8-level formats. The term *level* refers to the number of unit intervals or pulses in the intelligence-determining portion of the code. Each unit interval is either a *mark* or *space* as determined by the code for the desired character. The 5-level code has 2^5 (32) character permutations available and the 8-level code has 2^8 (256) available permutations.

To keep the transmitting and receiving machines in synchronization a start pulse

is placed in front of the group of intelligence pulses. A stop pulse is placed at the end of the group of intelligence pulses to complete the synchronization function. The start pulse is always a *space* condition and has the same pulse width or unit interval as an intelligence pulse. The stop pulse is always a *mark* condition and its minimum duration may be up to two unit intervals.

The 5-level code may be divided into three subcode types, depending on the width of the stop pulse. For example, a 60 word-per-minute 5-level code character includes the start pulse and five intelligence pulses, each of which has a pulse width of 22 milliseconds. Each 22-ms pulse or bit may be referred to as a *unit*. If the stop pulse in this group is also 22-ms wide then the group is called a 7-unit code. If the stop pulse is 31-ms wide then it is a 7.42-unit code. The 7.42-unit code is the most common 5-level code.

Another code in use is the 7.5-unit code where the stop pulse is 33-ms wide. The intended effect of the longer stop pulse is to decrease the amount of message garble under marginal operating conditions. However, the longer stop pulse has the undesirable effect of slightly decreasing the circuit speed capability.

functional description

The design objective was a simple, semi-programmable, all-electronic message generator using low-cost TTL IC logic packages and meeting the following requirements:

1. The required serial message format is: letters (LTRS), space, DE, space, K, X, figures (FIGS), 6, letters (LTRS), I, T, space, carriage return (CR), and line feed (LF).

2. The message generation cycle is initiated by an external momentary contact closure and/or a TTL compatible negative-going pulse.

3. The device must be self-stopping at the end of the message generation cycle.

4. The device keyer output must be compatible with any normal RTTY loop without regard to loop polarity or voltage level.

5. The device's message must be field programmable, either by means of plug-in boards or minor hardware changes, or both.

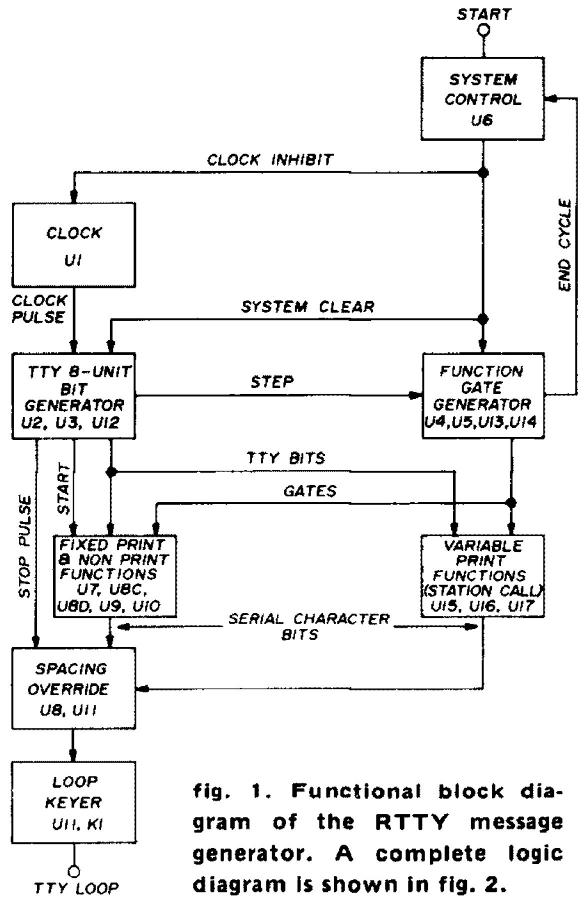
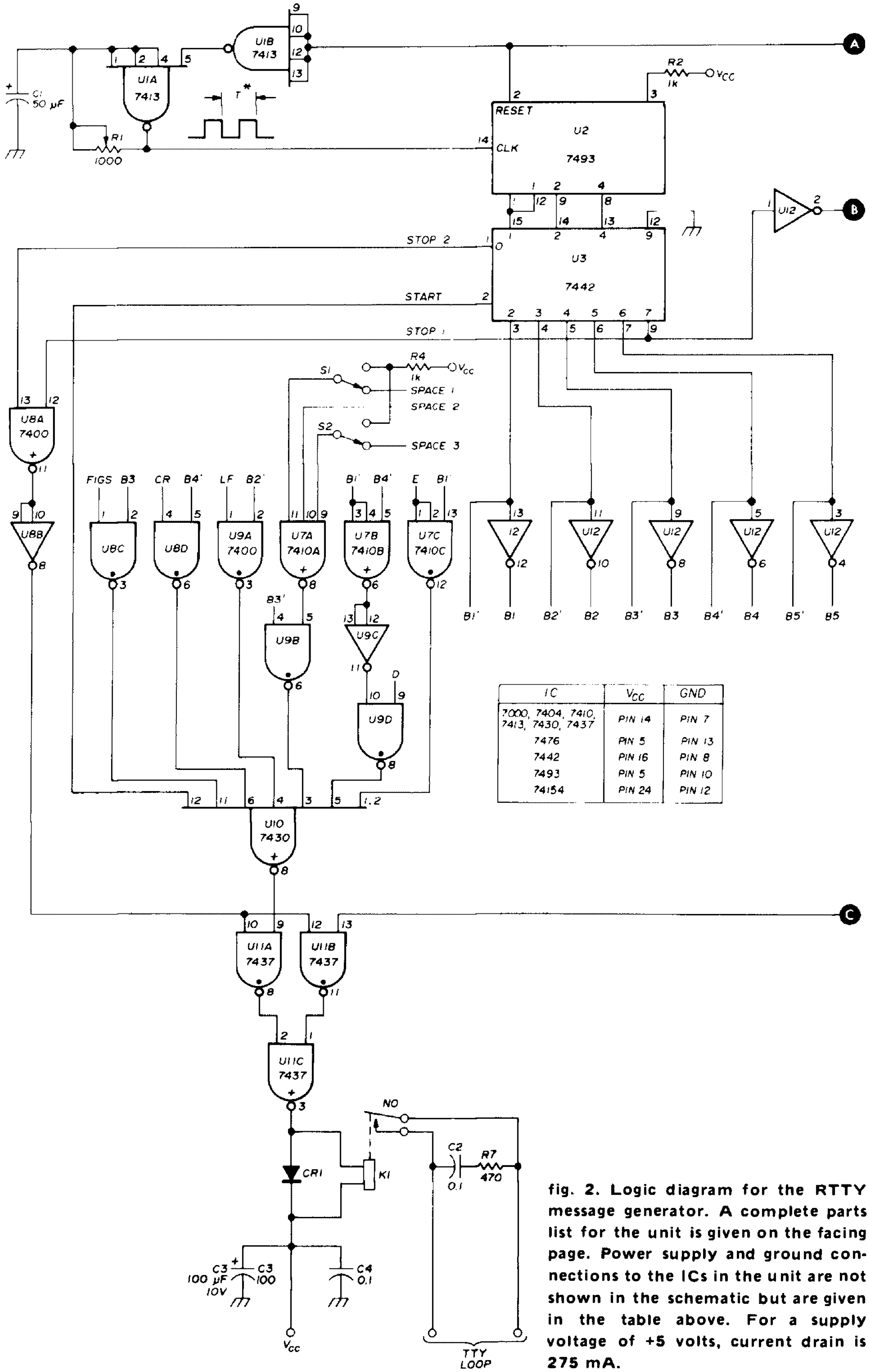


fig. 1. Functional block diagram of the RTTY message generator. A complete logic diagram is shown in fig. 2.

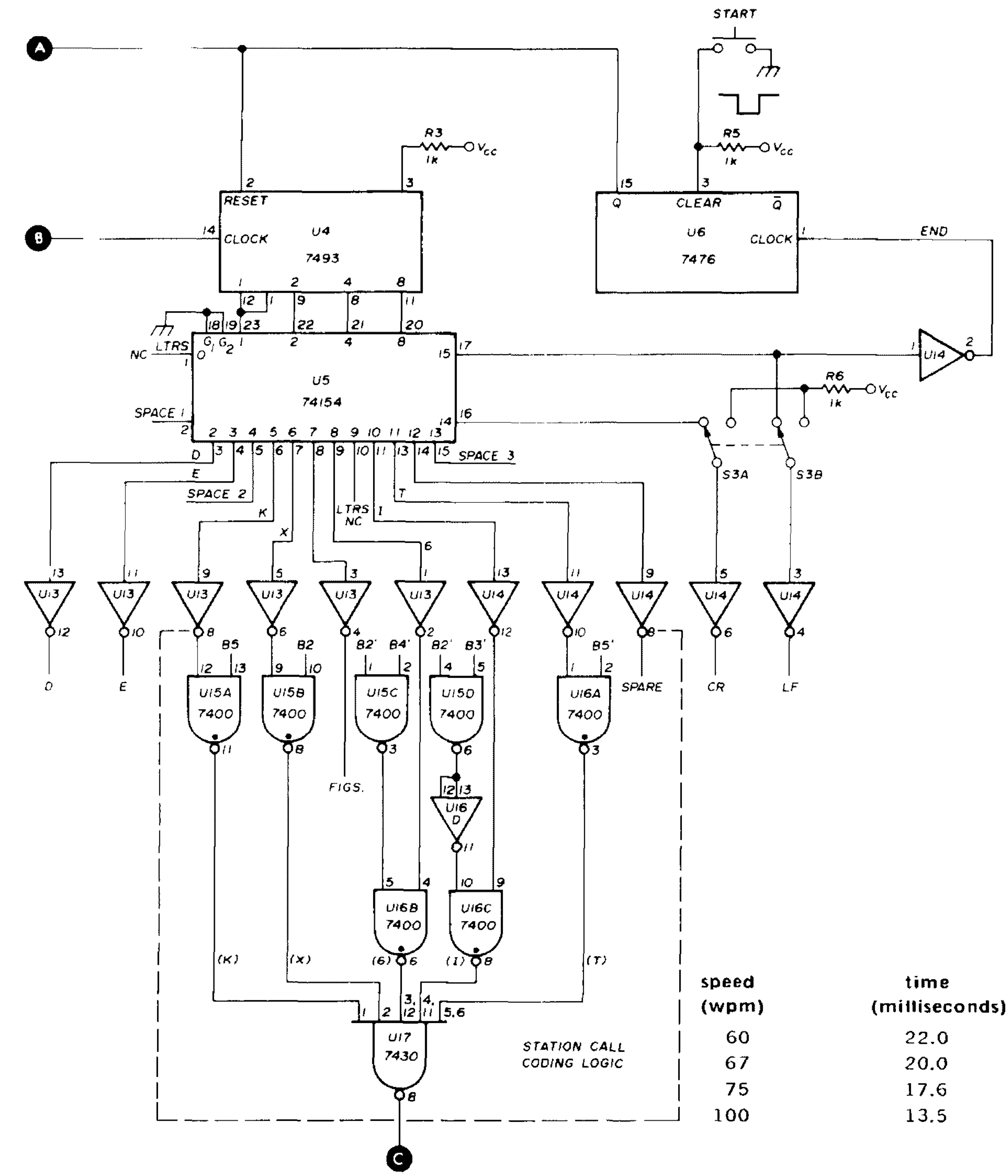
Fig. 1 illustrates the operation of the device at a basic functional block diagram level. A detailed logic diagram is shown in fig. 2.

System control. When the circuit is in an idle state, U6 generates a signal that inhibits 2 clock pulse generation (U1) and sets the 8-unit big generator and function gate generators to a cleared condition. On receipt of an external start signal, system control removes the clock inhibit and system clear signal. The device now begins the message generation cycle. At the end of the message, the function gate generator provides an end-of-cycle signal which returns system control to idle



IC	V_{CC}	GND
7000, 7404, 7410, 7413, 7430, 7437	PIN 14	PIN 7
7476	PIN 5	PIN 13
7442	PIN 16	PIN 8
7493	PIN 5	PIN 10
74154	PIN 24	PIN 12

fig. 2. Logic diagram for the RTTY message generator. A complete parts list for the unit is given on the facing page. Power supply and ground connections to the ICs in the unit are not shown in the schematic but are given in the table above. For a supply voltage of +5 volts, current drain is 275 mA.



speed (wpm)	time (milliseconds)
60	22.0
67	20.0
75	17.6
100	13.5

K1	spst printed-circuit relay (Clare LA-005, 5-volts, 380 ohms, DIP package)
K2	spst normally closed reed relay (Grisby-Barton GB821B-2)
S1,S2	spdt toggle switch
S3	dpdt toggle switch
U1	dual NAND Schmitt trigger (SN7413)
U2,U4	4-bit binary counter (SN7493)
U3	BCD-to-decimal decoder (SN7442)
U5	4-line to 16-line decoder/demultiplexer (SN74154)
U6	dual J-K master-slave flip-flop (SN7476)
U7	triple 3-input positive NAND gate (SN7410)
U8,U9 U15,U16	quadruple 2-input positive NAND gate (SN7400)
U10,U17	8-input positive NAND gate (SN7430)
U11	quadruple 2-input positive NAND buffer (SN7437)
U12,U13 U14	hex inverter (SN7404)

status, thereby terminating the message cycle.

Clock. IC U1 is connected as a gate-controlled pulse generator. The time between the negative-going edges of two adjacent pulses is set to equal the desired unit or bit width, i.e., 22 ms for a 5-level, 60-wpm machine.

RTTY 8-unit bit generator. The clock pulse from U1 is fed to the 4-bit binary counter, U2. The output of the binary counter is decoded by 1-of-10 decoder U3. This decoder sequentially produces eight unit bits each character generation cycle. In order of generation they are start, five intelligence bits and stop, which is 2 units in length. At the end of the 7th unit bit (halfway through the stop pulse) a step pulse is applied to the function gate generator. Complements of the bits are available through hex inverter U12. An 8-unit code is used instead of the standard 7 or 7.42-unit codes in the interest of circuit simplicity and minimum package count.

Function gate generator. The function gate generator is functionally similar to the 8-unit bit generator described above. The decoder section is a 1-of-16 decoder. The active function gate is advanced to the next decoded line each character generation cycle of the 8-unit bit generator. The last (16th) function gate pulse is inverted and applied to system control, U6, to terminate the message generation cycle. ICs U13 and U14 invert all function gates to match the character coding logic.

Fixed character. The 2-input and 3-input gate ICs in this block combine the active function gate and selected intelligence bits from the 8-unit bit generator to form the desired fixed print and non-print RTTY functions.

Variable character. This block is functionally similar to the fixed character block, combining function gates and selected bits to form the desired printing functions. It is labeled variable as this is the area of the circuit that can be programmed for different station call signs by use of plug-in circuit boards.

Spacing override. To realize gate and interconnection economy in the fixed and variable character circuits during the generation of certain characters, it was convenient to allow a spacing condition to exist at the outputs of these blocks during the stop-pulse generation period. The logic gates in the spacing override block ensure that the stop pulse is always fed to loop keyer, even if a spacing condition from the fixed or variable character blocks happens to be present simultaneously with the stop pulse.

Loop keyer. ICs U11C and U11D drive the loop keying relay, K1. Only one gate is used when driving a normally-open contact relay. The second gate is used as an inverter if a normally-closed contact relay is used. A high-voltage transistor could replace the relay if loop polarity is observed.

character coding logic

The idle condition of a teleprinter is the marking (loop current flowing) state. Moreover, examination of a coding chart reveals a slight preponderance of *mark* over *space* in the code as a whole if you disregard the seldom used *blank* character. Thus, it is logical to set up a condition at the loop keyer where it is only necessary to create a spacing condition at the proper intervals to generate the desired message.

The first space pulse in any character or machine function is the start pulse. In the letters (LTRS) function, where all five information pulses are marking, the start pulse is the only spacing pulse in the entire code group. Therefore, to generate a LTRS function, it is only necessary to apply the start pulse to the loop keyer – and the machine performs the LTRS function.

Refer to the logic diagram in **fig. 2** to follow the formation of the LTRS function. Initially, the circuit is in the standby state. Clock U1 is inhibited. Binary counters U2 and U4 are set to zero count.

One-of-ten decoder U3 is low on output zero and is high on the remaining 7 outputs (outputs 8 and 9 are not used for 5-level codes).

Output zero of U3 (pin 1) is labeled stop 2. This is the last half of the 2-unit stop pulse and is applied to U8A as a low level. The remaining input to U8A is a high level from output 7 (stop 1). The output of U8A is a high, inverted by U8B, and applied to both U11A and U11B as a low. Therefore, with one input of both U11A and U11B at a low level, the output of these AND gates will always be high, regardless of whether highs or lows appear at the remaining gate inputs.

For example, in the case of generating characters with only one or two information bits marking, it is convenient to set up the character coding logic so that a spacing condition (a high level at the output of U17) is applied to the remaining input of U11B during the last half of the stop pulse. Thus, a low on one input of U11B overrides the spacing condition, keeping the output of U11B high. This, in turn, keeps the loop in the marking state during the entire stop-pulse period.

Classification of characters according to number of marking pulses.

1	2	3	4	5
E (M1)	A (M1,2)	B (S2,3)	K (S5)	LTRS
T (M5)	D (M1,4)	C (S1,5)	Q (S4)	
SPACE (M3)	H (M3,5)	F (S2,5)	V (S1)	
CR (M4)	I (M2,3)	G (S1,3)	X (S2)	
LF (M2)	L (M2,5)	J (S3,5)	FIGS (S3)	
	N (M3,4)	M (S1,2)		
	O (M4,5)	P (S1,4)		
	R (M2,4)	U (S4,5)		
	S (M1,3)	W (S3,4)		
	Z (M1,5)	Y (S2,4)		

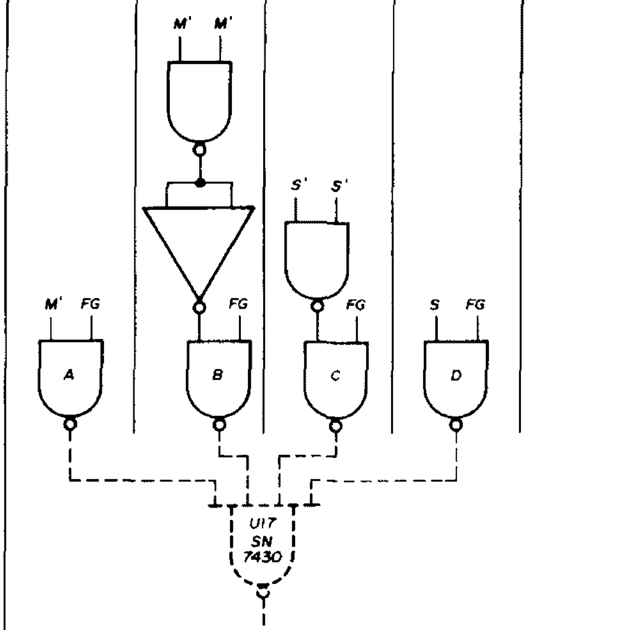


fig. 3. Callsign programming chart.

To initiate generation of the message and the first character (LTRS), momentarily depress the start switch, S4. This sets the Q output of flip-flop U6 to low, removing the inhibit from the clock, U1, and removing reset from U2 and U4. The first negative-going edge of the clock pulse toggles binary counter U2, causing output zero (stop 2) of U3 to go high and output 1 (start) of U3 to go low.

At this time both inputs of U8A are high, its output is low, and the output of U8B is now high and applied to one input of both U11A and U11B. Simultaneously, output 1 (start) of U3 is low and is applied to one input of U10, causing the output of U10 to go high. This high is applied to the remaining input of U11A. Both inputs of U11A are now high, causing the output to go low, creating a spacing condition at the loop keyer.

Thus, it may be seen that the loop is in a spacing condition immediately following arrival of the first negative-going edge of the clock waveform. It remains in this condition until the next negative-going edge of the clock again toggles binary counter U2; then output 1 (start) of U3 goes high and output 2 (intelligence bit 1) goes low. As soon as output 1 goes high, the output of U10 goes low, and the resulting high output of U11A causes the loop keyer to return to the marking condition. This sequence completes the generation of the start pulse, which is always a spacing condition.

Successive clock pulses applied to binary counter U2 move the low output of U3 through outputs 2 through 6 (intelligence bits 1 through 5). Since the function gate generator, U4 and U5, is still set

to zero, and because output zero of U5 (labeled LTRS) is not connected, no space pulses are generated during the periods of the five intelligence bits and the loop keyer remains in the marking state. Clock pulses continue to move the counter and decoder through 7 (stop 1)

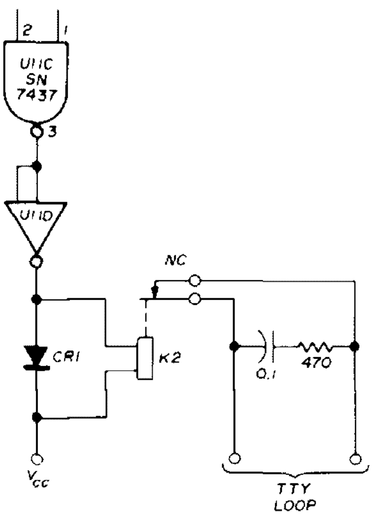


fig. 4. Alternate loop keyer circuit using a normally-closed relay. Relay K2 is a Grigsby-Barton 5-volt, 80-ohm spst relay in a DIP package. Diode CR1 is a silicon diode.

and returns it to output zero (stop 2). These two units of stop pulse complete the formation of the LTRS function.

space function

The machine space function character code has only bit 3 of the five intelligence bits in the *mark* state. Examination of 1-of-16 decoder U5 in fig. 2 shows that output zero (pin 1, labeled LTRS) is low during the idle condition and during the first character generation cycle of U3. At the time output 7 (stop 1) goes high, a pulse is applied to the clock input to U4. This changes the count from zero to 1 and moves the active low from output zero to output 1 in U5.

Output 1 from U5 (labeled *space 1*) is connected to one input of U7A through switch S1. This input goes low (all three inputs were high), the output goes high and is applied to U9B. At this time the B3 input to U9B is still high so the output goes low, causing the U10 output applied to U11A to go high. However, since the stop 2 bit applied to U8A is now low, the remaining input to IC-U11A, is also low, and, the loop keyer, U11C, continues to hold relay K1 in the marking state.

The next clock pulse applied to U2 moves the active low output of U3 to *start*. This low is applied to pin 12 of U10, but because of the low already on pin 3 of U10, the output and input to pin 9 of U11A remain high. At the same moment the active low in U3 moves from *stop 2* to *start*, the output of U8B goes high and the U11A output goes low, creating a spacing condition at K1 for the duration of the *start* pulse. Successive clock pulses continue to move the active low through the outputs of U3.

Because input to pin 3 of U10 remains in the high state, the loop keyer remains spacing throughout the periods of information bits 1 and 2. At the instant bit 3 goes low, the signal at pin 4 of U9B goes from high to low and U10 has all inputs high. This causes the loop keyer relay K1 to go to the marking condition for the duration of intelligence bit 3. Relay K1 returns to a spacing condition during the periods of bits 4 and 5 and then goes to marking during *stop 1* and *stop 2* periods. The *space* machine function character is now complete.

message characters

Completion of the *space* machine function character described above has advanced the count in U4 to three. Decoder U5 is now low on pin 3, labeled D. This low is inverted by one section of U13 and applied as a high level to pin 9 of U9D. The character D has intelligence bits 1 and 4 marking. These two bits are applied to the inputs of U7B. Both inputs are high at all times except during the periods of bits 1 and 4. Thus, a spacing condition exists at the output of U9D during the formation of the letter D except during the periods of bits 1 and 4, which are marking.

It is now apparent that as each character is completed, the gate function generator, U4 and U5, is advanced one count, and the associated active output is

applied to a logic gate or group of logic gates, enabling the appropriate selection of marking or spacing intelligence bits from the bit generator, U2 and U3, to form the desired characters.

Character generation continues until the beginning of the 17th pulse input to U4 which sets output 15 (pin 17) of U5 from low to high, and applies a negative-going level to the clock input (pin 1) of flip-flop U6. This causes the U6 Q output to go low, resetting both binary counters to zero and inhibiting the clock, U1, returning the message generator to idle. Should the clear input (pin 3 of U6) be held low continuously, it will override the end-of-cycle signal on pin 1 and the message generator will repeat itself until the low on pin 3 is removed.

programming

Switch S1 is provided to inhibit the *space 1* machine function if a space is not desired before the first printed character in the message. When *space 1* is inhibited, the message generator forms the non-printing machine function LTRS. Switch S2 inhibits a space after the last printed character in the message. Switch S3 inhibits the carriage return (CR) and line feed (LF) machine functions when a continuous line of print across the page is desired.

As many as four different character gating configurations are required for programming the generator. The gating configuration selected for a specific character is dependent upon the number of marking pulses in the character. **Fig. 3** tabulates characters according to their marking pulse content and illustrates the appropriate gating configuration. The notation FG at a gate input in **fig. 3** indicates connection to the inverted function gate originating at U5. The notation M′ indicates connection to the appropriate marking bit from the bit generator. Note that marking bits are selected only when the character contains one or two marking pulses.

The notation S′ indicates connection to the appropriate spacing bit from U3. Spacing bits are selected when the desired character contains three or four marking pulses. The numerals to the right of each character in columns one and two refer to the location of marking pulses in the 5-bit pattern. The numerals in columns three and four refer to the location of spacing pulses in the bit pattern.

In gate D (**fig. 3**) note the absence of a prime mark after the S input reference. This means that the spacing bit for characters in column four must be inverted instead of coming directly from the outputs of U3. Refer to connections in U8C and U12, pin 8 in **fig. 2** for an example.

As previously covered in the text, no gating or connections are required for the LTRS function.

construction

The physical configuration of the prototype message generator consists of two printed-circuit boards (main and station call) with edge connectors, a regulated power supply and a fully enclosed aluminum cabinet to provide radio frequency interference shielding as well as control mounting facilities. The main printed-circuit board is a universal dual in-line package (DIP) type breadboard with 15 sets of DIP IC pads for the 14 ICs and one DIP reed relay. Each IC pin pad has up to three solder pads for interconnection. The station call board is about half the size of the main circuit board and contains the three ICs indicated within the station call coding logic box in **fig. 2.**

Total cost of the IC packages for this unit is less than ten dollars. The cost of all components including ICs, power supply and transformer, but not including the printed-circuit boards, connectors and cabinet, amounts to less than $35.00. These costs are based on single unit prices.

Although not indicated in the logic diagram or in the parts list, the prototype unit uses a 4-position, single-pole rotary switch to select one of four 1000-ohm trimpots (R1) in the clock circuit. Each trimpot is adjusted for one of the four operating speeds listed in the speed-time chart in **fig. 2.** Also not shown on the

logic diagram are V_{cc}-to-ground bypass capacitors for ICs U1 through U6. These are 0.1-μF disc ceramic capacitors mounted as closely as possible to the V_{cc} and ground pins of each of the indicated ICs. These capacitors are required for suppressing noise generated by internal IC switching transients.

The spark suppression network (C2, R7) across the contacts of the keying relay is mandatory. Operation of the device without this network will result in premature failure of relay contacts, and in erratic operation of the circuit due to noise. Diode CR1 suppresses the voltage transient caused by back-emf generated in the coil of the relay at de-energization.

The normally open, spst reed relay, K1, is the type actually used in the prototype. It is less costly and easier to obtain from supply sources than the normally-closed, spdt reed relay, K2, shown in the alternate loop keyer configuration in **fig. 2**. Actually, the alternate configuration is preferred for most applications because loop continuity is maintained when power is removed from the unit.

The power supply consists of a 6.3 volt, one ampere power transformer and a rectifier-regulator circuit with 1% line-load regulation of the 5-volt dc V_{cc} output. The V_{cc} supply should be maintained within the limits of 5 volts, ± 5%.

troubleshooting

Troubleshooting improper operation is simplified if a typing reperforator is available as this permits recording of all normally non-printing machine functions on paper tape. Should a character not be the same as programmed, correlation of the tape readout with the appropriate area of the logic diagram should assist in isolating the problem. Experience has shown that almost all initial checkout problems in a handwired prototype result from improper or missing connections.

Rf interference can cause problems, although the prototype has functioned without error in the immediate vicinity of gain radiators with power inputs of 100 watts rms at 14 MHz. Most rfi problems can be cured with proper application of shielding and installation of bypass capacitors on all the input and output lines.

summary

A simple, reliable, low-cost method of generating short RTTY messages has been described. An operational prototype message generator using state-of-the-art integrated circuits has been constructed and tested under field operation conditions. This unit was built with components costing less than thirty-five dollars at unit quantity prices.

bibliography

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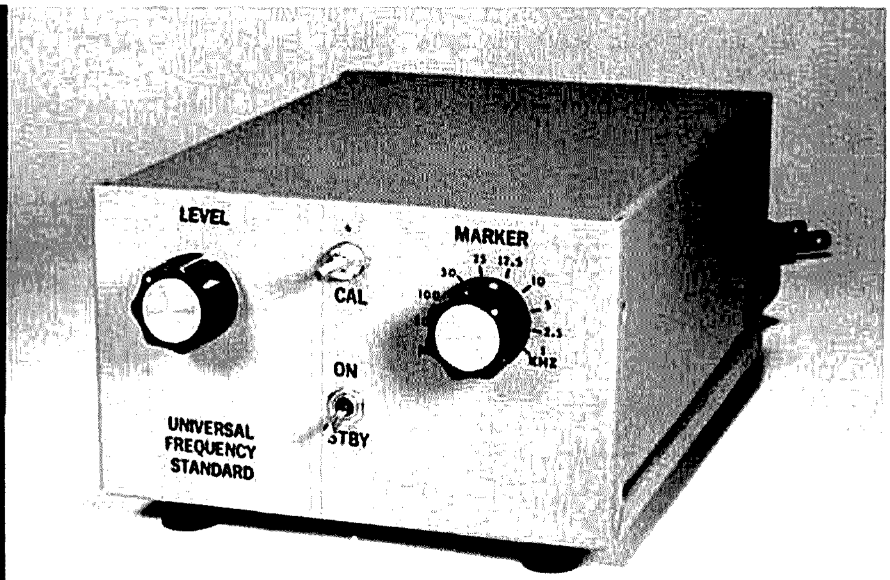
2. *Designing with TTL Integrated Circuits*, Texas Instruments Electronics Series, McGraw Hill Book Co., New York, 1971.
3. *The Integrated Circuits Catalog for Design Engineers*, Texas Instruments Incorporated, 1972.

ham radio

"I've discovered how to keep it playing a little longer at a time. I put a time-lag fuse in the box."



universal frequency standard

A precision frequency standard featuring a high-accuracy crystal, stable transistor oscillator and TTL logic

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

This frequency standard has been designed to supply precision frequencies for several purposes. Its main use is with my station receiver as a precise and reliable frequency calibrator. It may also be used as a digital counter time base, a scope calibrator or to drive a digital clock. With all integrated circuit packages installed on the board there are no less than eighteen frequencies available extending from 2 MHz down to 1 Hz. Ten of these, of your choice, are connected to a rotary switch for calibrator use while the lower frequencies below 1 kHz that would not normally be used with the station receiver, are picked off terminals on the circuit board by a small clip lead when needed.

The design is flexible. If you do not need the versatility of the complete unit, the photo shows a frequency standard that will provide markers at 1 and 2 MHz, and at 500, 250, 100, 50, 25 and 12.5 kHz with the installation of only three digital IC packages. Provision is made on the circuit board for as many or as few output frequencies as are likely to be needed.

circuit features

Features include excellent frequency stability, front panel calibration and a precise, self-contained, regulated power supply. An adjustable level control is included so the calibrator output can be matched to incoming signals such as WWV for really accurate zeroing or advanced full on for strong, clear markers.

With proper temperature compensation, the frequency standard will stay within 1 Hz of WWV at 10 MHz over an extended period of time with no adjustment.

This precision is not needed if the unit is used only to find band edges or set the receiver graticule. However, for frequency measurement or use as a time base for a counter or digital clock, you need all the accuracy you can get. The best reason for going first class is that it costs little more, and probably less, in this do-it-yourself project. The required stability can be obtained with inexpensive construction.

The oscillator circuit in this calibrator was first used as a stable time base for a digital clock and later in a receiver.[1] Several circuits evaluated in drift tests showed the Clapp-Colpitts to be most stable — not a new circuit, but seldom used in recent designs. While both the circuit used and the TTL logic is familiar, it is the combination of circuit, crystal and construction that makes this calibrator better. Stability comes when a few hertz drift is removed or greatly reduced from each of several sources. A 10-Hz drift caused by a trimmer might be tolerated, but when the drift from other sources is combined, the total becomes excessive.

Most oscillator circuits would cause even a perfect crystal to drift. Voltage and temperature changes cause changes in the semiconductor's internal impedance. Trimmers don't stay where set, wires move, capacitors change value. Solid-state circuitry has substantial advantages over vacuum-tube circuits; much heat is eliminated, components run cooler and the crystal is driven at the low levels recommended by the manufacturer. However, a substantial amount of drift can come from the semiconductor alone.

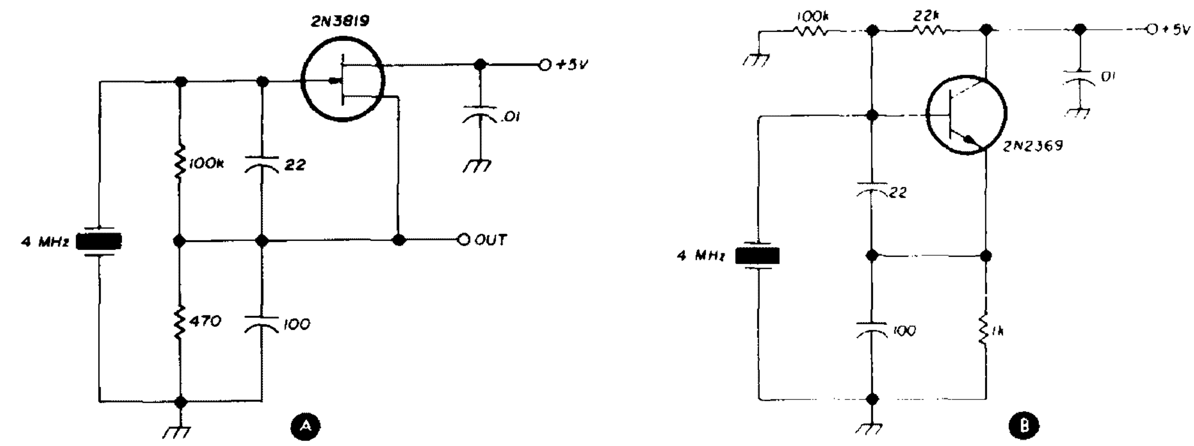


fig. 1. Solid-state crystal oscillator circuits. Fet circuit in (A) is considerably more stable than the bipolar circuit shown in (B). For a more stable bipolar oscillator circuit, see fig. 2.

How much? Two circuits are shown. **Fig. 1A** shows a fet oscillator circuit using crystals in the 1- to 9-MHz range. The crystal ground for 32-pF load is sometimes brought on frequency with a small trimmer. If this circuit is built so the semiconductor is isolated, it can be heat cycled without affecting other components and the drift can be measured on a

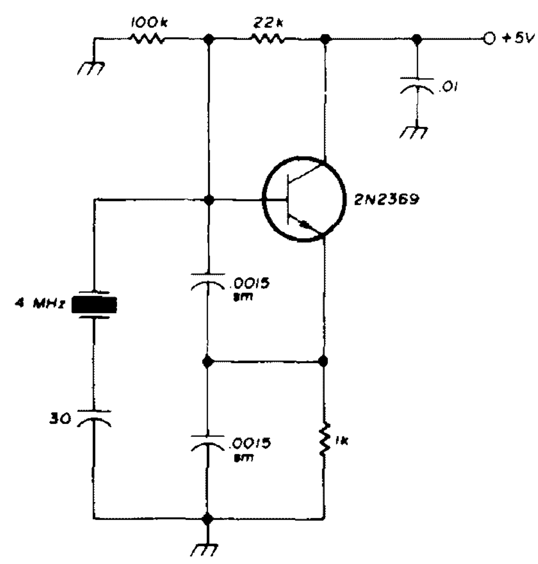
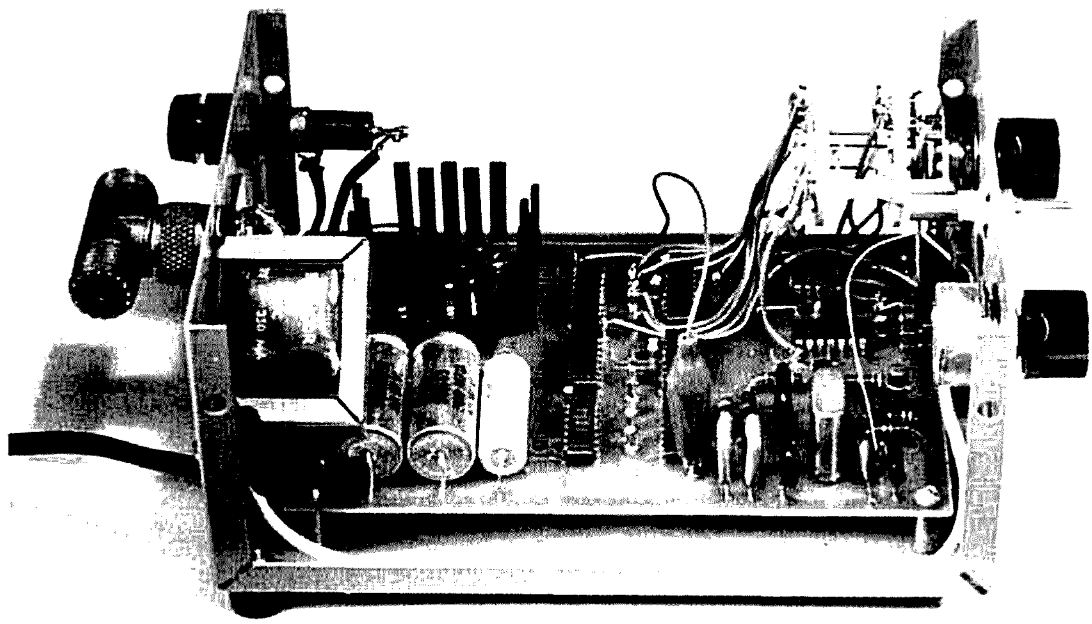


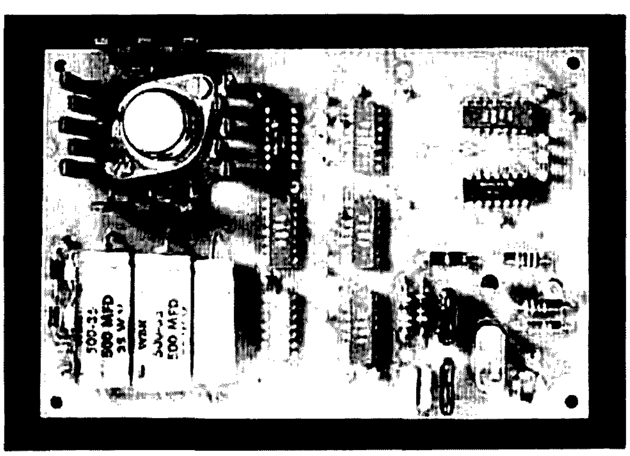
fig. 2. This transistor crystal oscillator circuit is stable because impedance changes are swamped out by the 1500-pF silver-mica capacitors.



Frequency standard and power supply are installed in a 3x5x7-inch (7.6x12.7x17.8-cm) Minibox.

digital counter. The best of several fets caused a change of 12 Hz at 4 MHz with a 5-degree ambient temperature variation.

Fig. 1B shows a transistor in a similar oscillator circuit, not recommended, but sometimes used. Although not as sensitive as the fet to temperature, a change of only one volt caused a 70-Hz frequency shift. If this circuit is modified as shown in fig. 2, there is a substantial improvement. Changes of ten degrees and one volt did not cause a frequency change readable on the counter. The transistor case could be heated with a soldering iron to the point where it burned the fingers with a 2 Hz change registered at 4 MHz.



Printed-circuit board with all logic packages installed, as would be required for driving a digital clock or a frequency counter. One jumper is used.

These experiments show both the extent of drift that can be contributed by the semiconductor and indicates the solution. The more stable the capacitance used across the transistor, the better the stability. The 1500-pF capacitors have a very low reactance at 4 MHz, swamping out any other impedance changes in the circuit. This circuit will not oscillate with some transistors because there must be enough gain to sustain oscillation. This requirement is met by the Motorola HEP715, a pnp device with a typical beta of 120. Other transistors with similar current gain can be substituted.

the crystal

Some time ago while working with digital counter time bases[2] I noticed that the ordinary 100-kHz crystal was not as stable as it might be. A time base derived from the power line was nearly as accurate. Although used in amateur calibrators for many years, the 100-kHz rock is

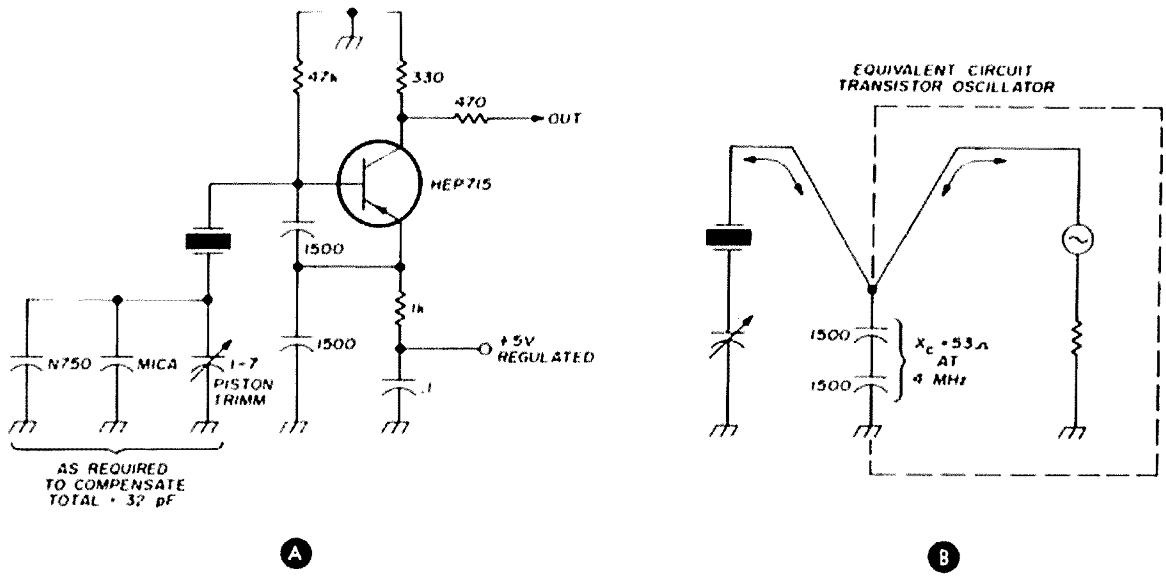


fig. 3. Stable transistor crystal oscillator circuit. Equivalent circuit in (B) shows how 1500-pF capacitors swamp out any internal impedance changes which would cause the output frequency to drift.

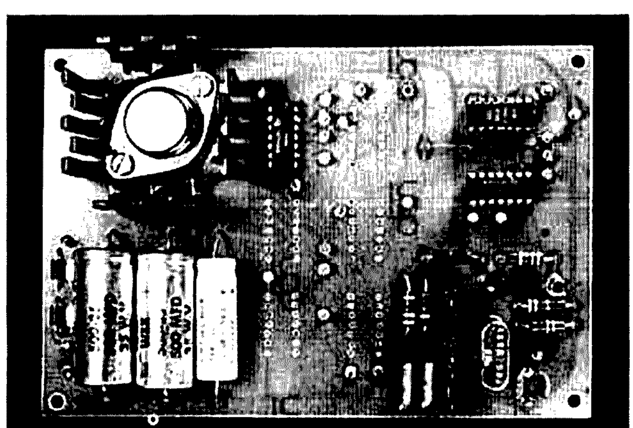
not a good choice as it must be stabilized by use of an oven; this adds bulk, expense and a heater supply.

A better crystal for a frequency standard is the high-accuracy 4-MHz crystal recommended by W6FFC.[3] High accuracy as used here means the crystal drifts less, and at a predictable rate. Although cheaper crystals in the 1- to 5-MHz range are better than those at 100 kHz, they are subject to random drift which is difficult to compensate. Therefore, it is important to obtain the better quality crystal. They are manufactured by both Sentry and International and cost less than $10.*

Note the drift characteristic curves for a typical AT-cut crystal in fig. 4. As the temperature increases, the crystal frequency decreases. Drift can be almost entirely eliminated at room-temperature operation which most amateurs are interested in by selecting the proper value of negative coefficient compensation capacitor. Curve A requires the most compensation while curve B requires little if any. A crystal cut for a 32-pF load allows about 30 pF for compensation with the piston trimmer adding the 2 or 3 pF needed to pull the crystal to the exact frequency.

*Write for their catalogs. Sentry Manufacturing Company, Crystal Park, Chickasha, Oklahoma 73018; International Crystal Manufacturing Company, 10 North Lee, Oklahoma City, Oklahoma 73102.

A new crystal might require anything from a maximum of 30-pF (N1500) to a 30-pF silver mica (NPO), depending on its temperature vs frequency characteristic. Since there is no way to know what will be needed, it is advisable to have a supply of different values of N750 and N1500 coefficient capacitors on hand, with small silver micas to pad the total to 30 pF, before starting any temperature compensation work. A piston trimmer is recommended because of the smoother adjustment and lack of drift. It is also easier to determine if capacitance is being added or removed, useful information when temperature compensating the calibrator.



If you don't need the versatility of the complete unit, this photograph shows a frequency standard that will provide markers at 1 and 2 MHz, and at 500, 250, 100, 50, 25 and 12.5 kHz with the installation of only three logic packages.

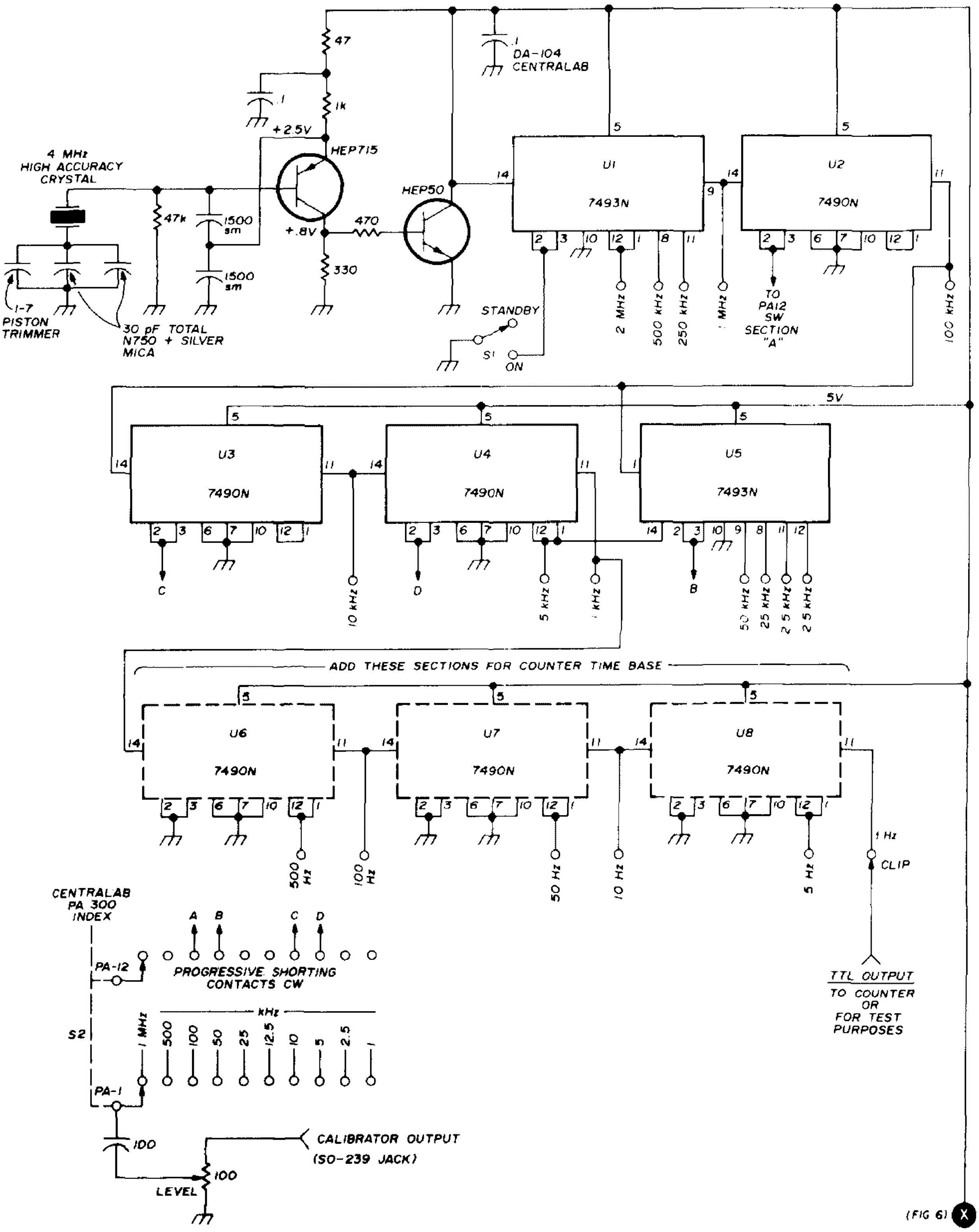


fig. 5. Complete schematic diagram for the universal secondary frequency standard. Unit uses high performance TTL logic ICs. A printed-circuit layout is shown in fig. 6.

the circuit

In fig. 5 the pnp oscillator transistor, Q1, is coupled to the TTL logic by transistor Q2. The 7493 binary dividers U1 and U5 are used to divide by factors of 2, with 7490 decade packages making up the remainder of the logic. IC U5 has two inputs for 5 kHz and 100 kHz. Reset pins 2 and 3 control operation of the logic, either by switch S1, the standby switch, or by progressively shorting contacts on the rotary switch.

This way, the oscillator runs contin-

uously for best stability, and unused packages are disabled. It prevents some markers from leaking across the selector switch and being heard in the receiver. If this feature is not wanted, pins 2 and 3 should be jumpered to ground.

Board outputs and compensating capacitor terminals appear at convenient terminals at the top of the board made by forcing short lengths of bare number-12 wire into 5/64-inch holes. This facilitates exchange of compensating capacitors, or selection of a different logic output at some future time. After completion, it is difficult to work on the underside of the board without removing several wires.

The fast switching TTL logic has active transistor pull-up circuitry which is well suited to driving external loads. The 2900th harmonic of the 10-kHz marker is over S9 in the ten-meter band. In the unlikely event that a logic package fails, repair would be facilitated if Molex sockets are used.

The full current drain with all IC packages installed is 5 volts at 260 mA, easily supplied by a LM309K voltage regulator IC mounted on a heatsink (fig. 6). All power supply components except the power transformer are mounted on the circuit board. High temperature shutdown and overcurrent protection are provided by the regulator.

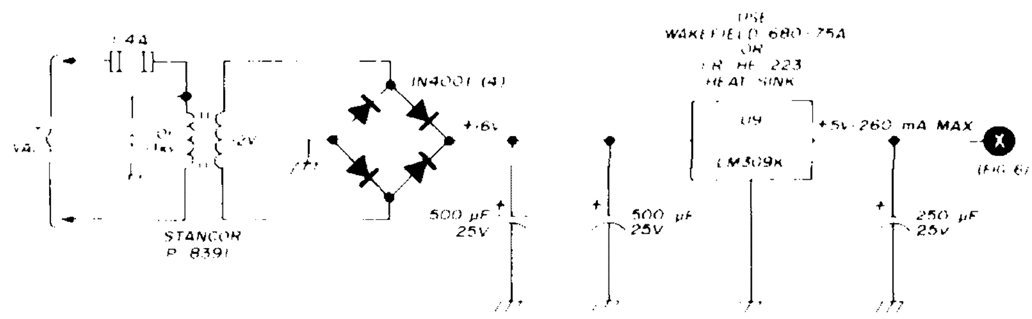


fig. 6. Power supply for the universal frequency standard.

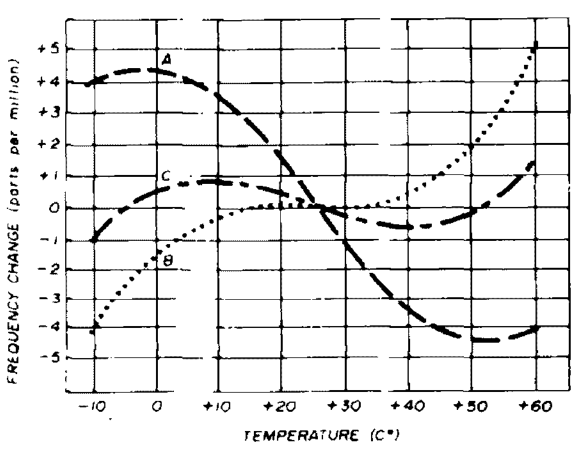


fig. 4. Frequency vs temperature chart for typical high-accuracy AT-cut crystals.

construction

This frequency calibrator is simple to build. The circuit board may be hand duplicated following the layout given in fig. 7, or an etched, plated epoxy board is available which speeds construction and minimizes errors.* Parts locations and identifications are screened on the board. It is only necessary to drill the IC holes with a number-60 drill, insert parts, and solder. The assembly is mounted in a compact 3x5x7-inch Minibox.

temperature compensation

Crystals should be ordered for 0.0005% tolerance, F-700 or SP7-P holder (depending on manufacturer), 32-pF load, 4 MHz at room temperature. New crystals should be operated for a time before starting any compensation work. You will need the previously mentioned supply of N750 and N1500 capacitors, and a receiver with an S-meter that will tune WWV.

Start with 15 pF in parallel with a 15-pF N750, allow an hour for the unit to stabilize, and adjust to frequency using the S-meter on the receiver as an aid to exact zeroing. Select a time when WWV is

*An epoxy, plated 4x6-inch printed-circuit board for this frequency standard is available from the author. $8.00, postpaid, in the United States.

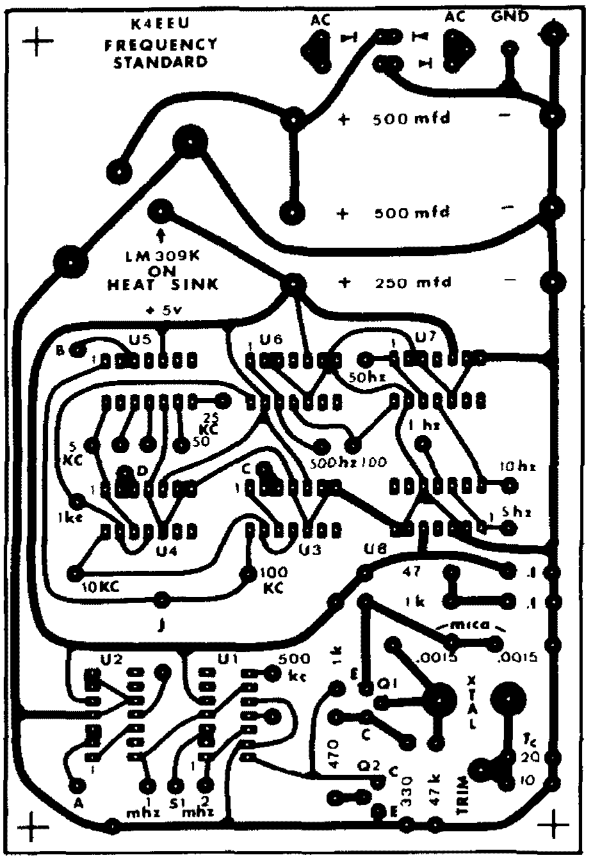


fig. 7. Printed circuit layout for the secondary frequency standard. Boards are available from the author for $8.00, postpaid.

moderately strong, with little fading, so the meter remains reasonably steady.

The standby switch, S1, is turned on and the calibrator *level* control advanced about halfway so both signals are heard. At first, the calibrator will probably be so far off frequency that an audible beat note will be noted, mixed with the WWV tone. As it is zeroed, the warble in the WWV tone will decrease in pitch until it is no longer heard and the S-meter will swing, rapidly at first, then slower, as tuning becomes more exact.

The amplitude of this swing will maximize when the calibrator level is equal to WWV's strength. It should be easy to set the calibrator exactly in zero beat with WWV at 10 or 15 MHz. When this is finished, note the temperature of the room on a thermometer, and record it for reference. Recheck the frequency with WWV periodically, and if there is any drift note the temperature and the direction the trimmer must be adjusted – to add or remove capacitance. If this trimmer capacitance must be reduced for a temperature increase, more N750 or even N1500 capacitance is needed, always padding the combination to a total of 30 pF. after a few tries the exact value of compensation will be found.

Take your time during this work and be sure a trend is established before changing capacitors, possibly making two or more observations before proceeding. The temperature compensation is easier to accomplish than it appears and makes the difference between an ordinary and a precision instrument.

some uses

The photos show a frequency standard supplying pulses for a digital clock. This clock controls nineteen slave clocks in a broadcast installation, so reliability and accuracy are important. The clock is made immune to momentary power-line failures by floating the dc supply across a nicad battery large enough to operate the logic until emergency power can be started or service is restored, whichever comes first.

Fig. 8 shows diodes used to drop the battery voltage to TTL requirements. The one-second pulses from this standard are so accurate that this clock stays on the tick with WWV for weeks with no correction.

In this circuit the 4-MHz crystal frequency is divided by a factor of 4 X 10^6 to obtain the one-second pulses. If this crystal drifted an extreme 1 Hz away from nominal, it would require four million seconds (or 46.3 days) for the clock to accumulate an error of one

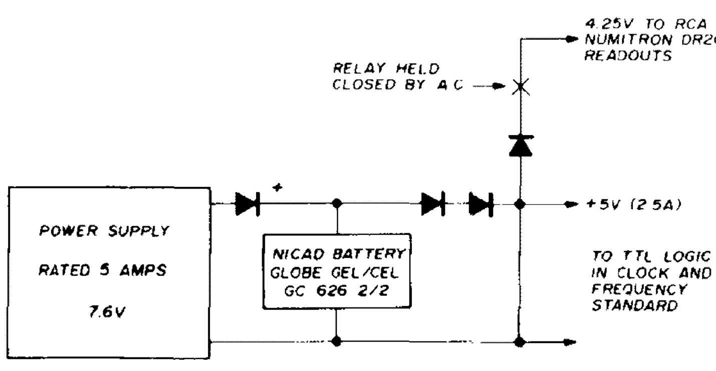
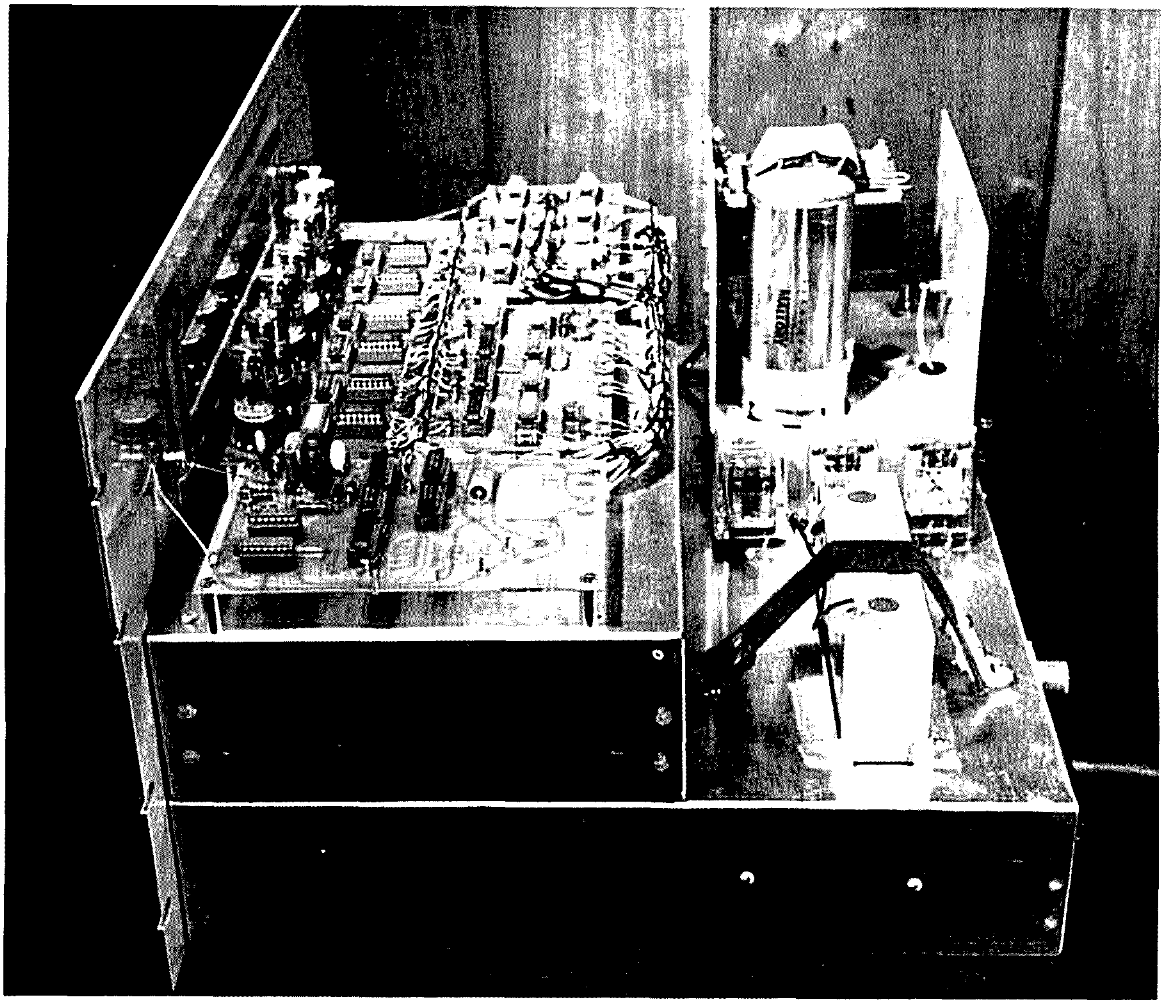


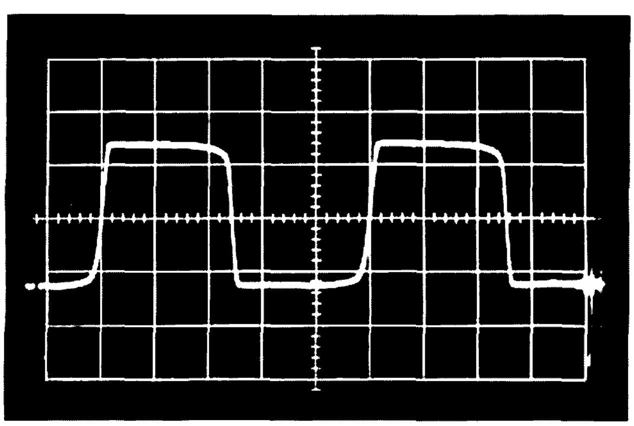
fig. 8. Failsafe power supply used with the digital clock described in the text.



Precision digital clock discussed in the text. The frequency standard board used to provide the 1-second driving pulses is located in the foreground on the upper deck. Nicad battery to the right is part of the failsafe power supply (see fig. 7).

second. But, since the crystal is compensated closer than this, and any minor drift is above and below the frequency, the average error is very small.

The secondary frequency standard would also make a good time base for a digital or Rec-Counter.[5] Readout kits with the tubes, storage latches and counter ICs are advertised in this magazine, so only the gating circuitry would have to be hand wired. Other applications for the standard include audio oscillator or signal generator calibration, or calibration of the sweep time base in oscilloscopes.

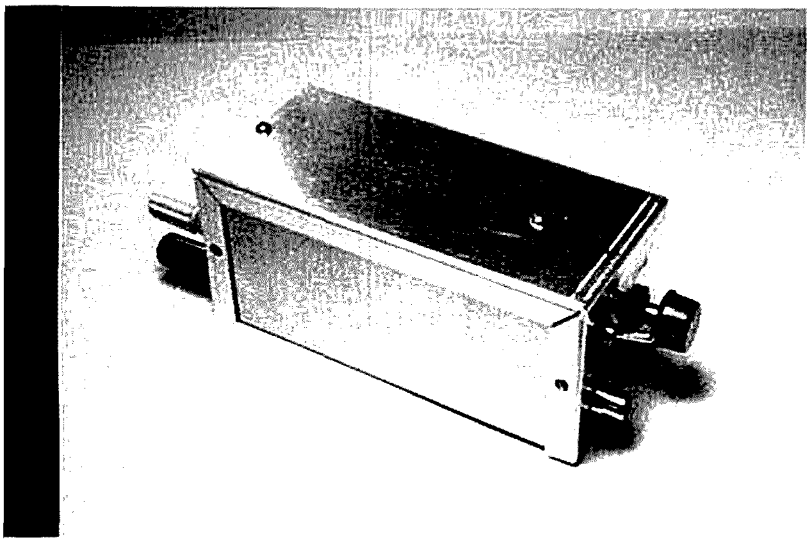


1-MHz output of the frequency standard as observed on a 10-MHz oscilloscope. Rounded waveform shows bandpass limitation of the scope. Horizontal scale is 0.2 microsecond per cm.

references

1. A.A. Kelley, K4EEU, "RTTY Monitoring Receiver," *ham radio,* December, 1972, page 27.
2. A.A. Kelley, K4EEU, "15-MHz Digital Frequency Counter," *ham radio,* December, 1968, page 8.
3. Irvin M. Hoff, W6FFC, "The Mainline FS-1 Secondary Frequency Standard," *QST,* November, 1968, page 34.
4. A.A. Kelley, K4EEU, "How to Make your Own Printed-Circuit Boards," *ham radio,* April, 1973, page 58.
5. Kenneth Macleish, W1EO, "A Frequency Counter for the Amateur Station," *QST,* October, 1970, page 15.

ham radio



455-kHz i-f alignment signal generator

Simple, crystal-controlled signal generator for aligning i-f strips — modulation is built in

Courtney Hall, WA5SNZ, Dallas, Texas 75240

I recently completed a homebrew receiver project which included a 455-kHz i-f strip, and I needed a signal source to align it. My home workshop doesn't boast a signal generator, and even if I had one, I would have no way of precisely setting its output frequency to 455 kHz. A little thought and investigation provided a fairly cheap and easy solution.

circuit

A schematic of the resulting generator is shown in fig. 1. It consists of an fet crystal oscillator and an amplitude modulator. I decided on crystal frequency control because it would set the frequency accurately and because 455-kHz crystals are available from JAN Crystals for $1.75 plus 10 cents postage.* These crystals are supplied in an FT-241 holder;

*JAN Crystals, 2400 Crystal Drive, Ft. Myers, Florida 33901.

the pins are 0.093 in diameter with 0.486 spacing. JAN sells the mating SSO-1 socket for 15 cents.

Transformer T1, the drain load for the fet oscillator, is a 455-kHz i-f transformer salvaged from a junked a-m transistor radio. This provides a simple way to tune

A Colpitts audio oscillator is used to provide amplitude modulation, and a switch allows the modulation to be turned on or off. A surplus 88-mH toroid is used in the audio oscillator. These can be found listed in surplus and classified ham ads for about 50 cents each or less.

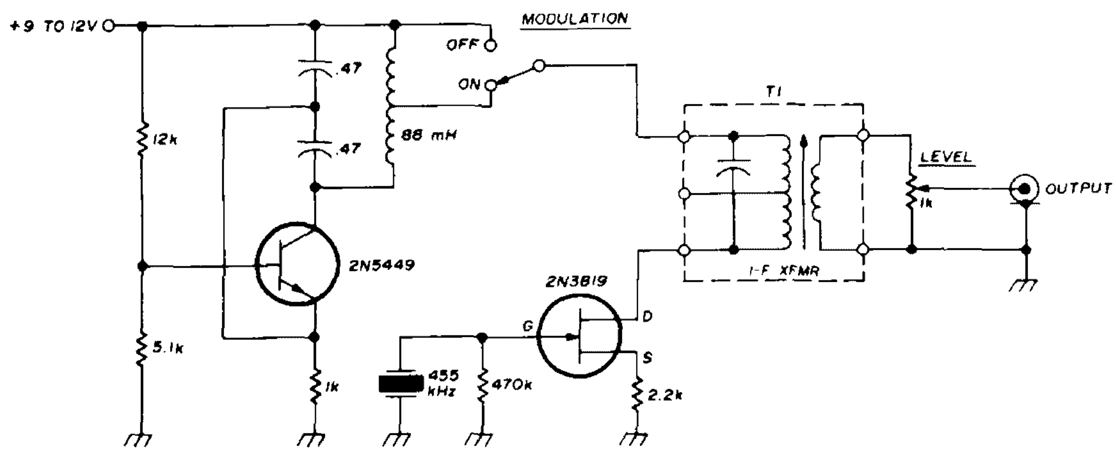


fig. 1. Schematic for the 455-kHz signal generator. Transformer T1 was salvaged from a junked transistor broadcast radio.

the drain circuit and obtain a low output impedance. Removing these transformers from the radio PC board is tricky because you have to simultaneously melt the solder at several different points. A large soldering iron is an advantage.

The junked radio had three i-f transformers, and I tried all three. I couldn't get the first one (mixer output) to oscillate at all, but it may have been damaged in removing it from the PC board. The last i-f transformer (which feeds the detector) had the highest output, but oscillations stopped if it was loaded with less than 200 ohms. I used the middle transformer because it would still oscillate when loaded with 50 ohms.

Another source of i-f transformers is Radio Shack. They sell a kit of four transformers for $1.39 (catalog number 273-1383). I believe the one in this kit which would correspond to the one I used is color-coded white.

The one I used has four leads, and two adjacent leads must be tied together to provide the center-tap. The audio frequency is about 1-kHz, but this may be altered by changing the value of the 0.47-μF tank capacitors.

I used a 2N3819 fet, but a Motorola MPF102 or Siliconix U183 should perform identically. The 2N5449 modulator should be available at local Radio Shack stores for 79 cents (catalog number 276-2014).

construction

As shown in the photographs, the generator is housed in a 2¼x2¼x5-inch Minibox. A piece of perfboard holds most

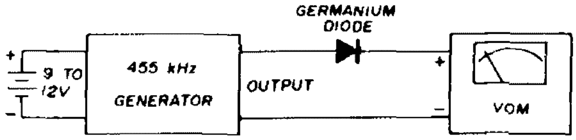
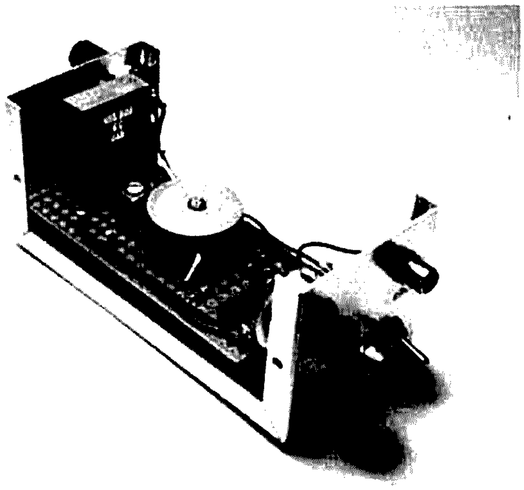


fig. 2. Arrangement used to adjust transformer T1 for maximum rf output.



Layout of the 455-kHz i-f signal generator. All components are wired on a section of perfboard which is mounted in a small Minibox.

of the circuit components. It is mounted in the Minibox by two screws with ½-inch spacers. The 88-mH toroid is held to the perfboard by a screw and two discs (one metal, one plastic) which were furnished with the toroid. I detected nothing critical in the layout.

One end of the Minibox holds the modulation switch, output phono jack and level control. Two 5-way binding posts are mounted on the other end for connecting the generator to a dc power source.

operation

The fet may not oscillate until T1 is adjusted. Connect a sensitive vom to the output through a germanium diode detector as shown in fig. **2**. Set the level control at maximum and the vom to its most sensitive dc volts scale. Now adjust the tuning slug in T1 for a maximum reading on the vom. This will only be a fraction of a volt, but this is more than enough for i-f amplifier alignment. The level control pot will not set the output voltage low enough for sensitive i-f circuits, and I found it necessary to use the external attenuator shown in fig. **3** to prevent overdriving the i-f strip.

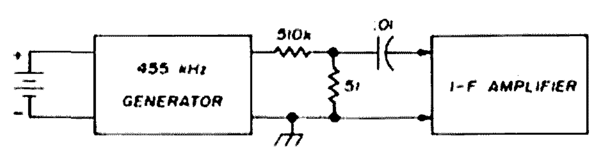


fig. 3. External attenuator which may be used to decrease generator output.

Although the circuit was originally designed to operate from a 12-volt dc supply, it appears to perform well using only a 9-volt transistor radio battery for a power source. Current drain is about 7 mA with a 12-volt supply and 5 mA using the 9-volt battery.

conclusion

This little gadget is intended only for 455-kHz i-f alignment which is a rather limited use. However, it has a limited cost too – only a few dollars. A well stocked junk box can cut the dollar outlay to a very nominal amount. I haven't had an opportunity to check its frequency on a counter or observe its output waveform on an oscilloscope, but it performed its intended function to my satisfaction. If amplitude modulation is not required, the unit could be simplified to a single fet circuit. This would reduce battery drain substantially. A worthwhile addition would be a built-in step attenuator which would permit setting the output voltage to micro-volt levels.

ham radio

"The wife just gave me a hint of something I need to do!"

multichannel fm receiver for six and two

How to use commercial fm receiver strips in a multichannel, two-band vhf fm receiver

Stirling Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

Surplus public-service vhf-fm equipment, which is sold in *strip* form, can be used as the basis for a low-cost, multichannel, two-band fm receiver. This is accomplished simply by adding a two-meter converter and a logic crystal oscillator[1] to a single-channel 30-50 MHz Motorola Sensicon receiver strip as shown in block form in fig. 1. If the output of the two-meter converter is fed into the strip receiver's first-conversion i-f input, and a logic oscillator is provided as the local oscillator (17.775 to 17.830 MHz) the strip will cover the fm channels for 146.34/94 MHz. By adding a multiple crystal oscillator operating at approximately 16 MHz and appropriate switching controls, two-band control and channel selection is possible.

These modifications are not limited to Motorola receivers, as there are a number of commercial fm receiver strips sold in the same way. Any of these strips can be adapted to perform the same task. However, before modifying one of these units it's a good idea to put it into operating condition before adding to the confusion. A circuit diagram and receiver tuneup data, if you can find them, are a great help in this respect. Information on many of these units is included in *The FM Schematic Digest.*[2]

two-meter converter

If you already have a good two-meter converter, all you have to do is convert its output frequency to the same frequency as the receiver strip you are going to use (4.3 MHz in the Motorola Sensicon receiver strip). This may be as simple as

plugging in the two-meter logic oscillator and realigning the converter, or it may require more extensive circuit modifications. If considerable modification is required, it might be easier to build a two-meter converter specifically for use with the fm receiver strip.

My two-meter converter consists of a single mosfet rf stage using an RCA 40822 mosfet. Another mosfet, an RCA 40823, is used as the mixer (see fig. 2). Both of these devices were designed for vhf work and provide good performance on two meters. The rf amplifier has excellent gain as an unneutralized rf amplifier, a low noise figure, and wide dynamic range which results in low cross modulation. The dual-gate mosfet used in the mixer stage isolates the input from the output and allows low-level local-oscillator injection.

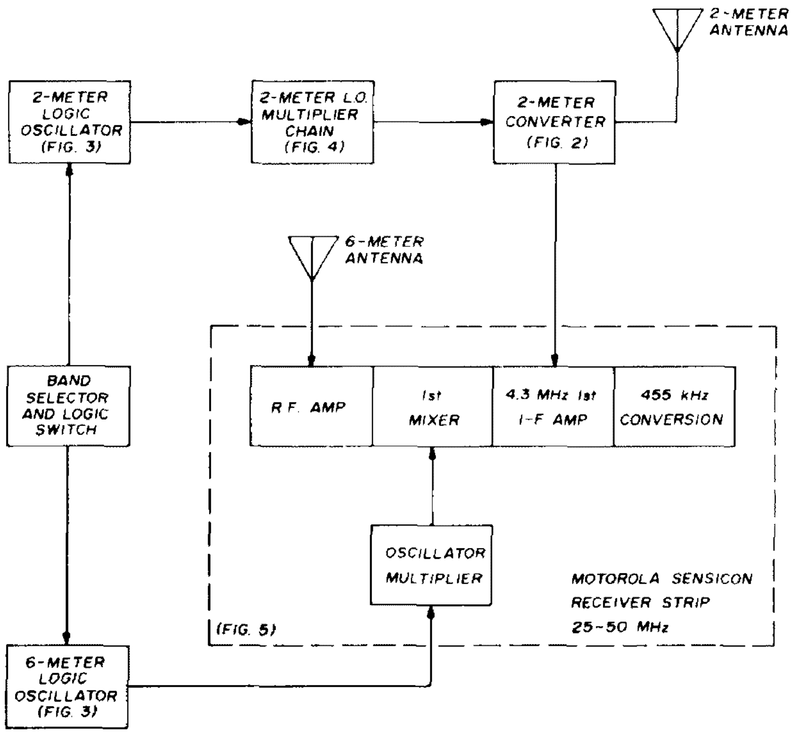


fig. 1. Block diagram of the multichannel six- and two-meter fm receiver using a Motorola Sensicon receiver strip.

The tuned input network to the rf amplifier is designed to match a 50-ohm antenna. The small trimmers, C1 and C2, the inductor, L1, and the rf amplifier transistor, Q1, are located in a shielded compartment made from 1-inch-wide strips of copper-clad PC material. The drain lead of Q1 passes through a small hole in the shield wall and is connected to inductor L2. Gate 2 and the source lead of the mosfet are connected directly to 1000-pF standoff capacitors. The 275-ohm source resistor is grounded next to the source bypass capacitor with as short leads as possible. A ferrite bead is installed on the drain lead.

Similar construction is used for the mixer stage. A small coaxial cable must be used to connect gate 2 of the mixer to the output of the multiplier chain. This is because the mixer requires only a small amount of local-oscillator signal — unwanted signals can leak in and appear in the 4.3-MHz output.

local oscillator

Construction of the logic oscillator will not be discussed as that was covered in detail in the previous article.[1] Crystal frequencies for the logic oscillator may be determined from the following formula

$$f_{xtal} = \frac{f_o - f_{i\text{-}f}}{8} \qquad (144\text{ MHz})$$

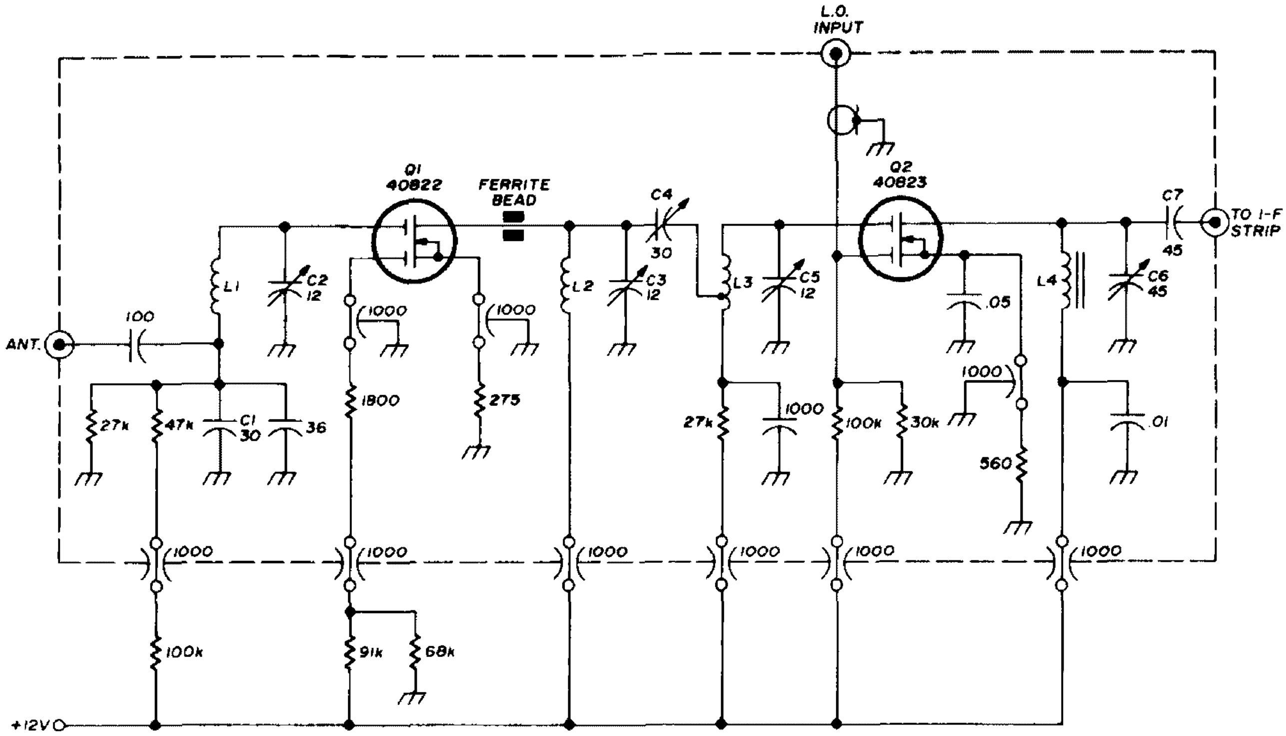
$$f_{xtal} = \frac{f_o - f_{i\text{-}f}}{3} \qquad (50\text{ MHz})$$

where f_{xtal} is the crystal frequency, f_o is the desired operating frequency and $f_{i\text{-}f}$ is the intermediate frequency of the receiver strip (often 4.3 MHz). For example, the crystal required for a two-meter input frequency of 145.500 MHz is

$$f_{xtal} = \frac{145.500 - 4.30}{8} = 17.650\text{ MHz}$$

For a six-meter input at 52.525 MHz, the required crystal frequency is

$$f_{xtal} = \frac{52.525 - 4.30}{3} = 16.075\text{ MHz}$$

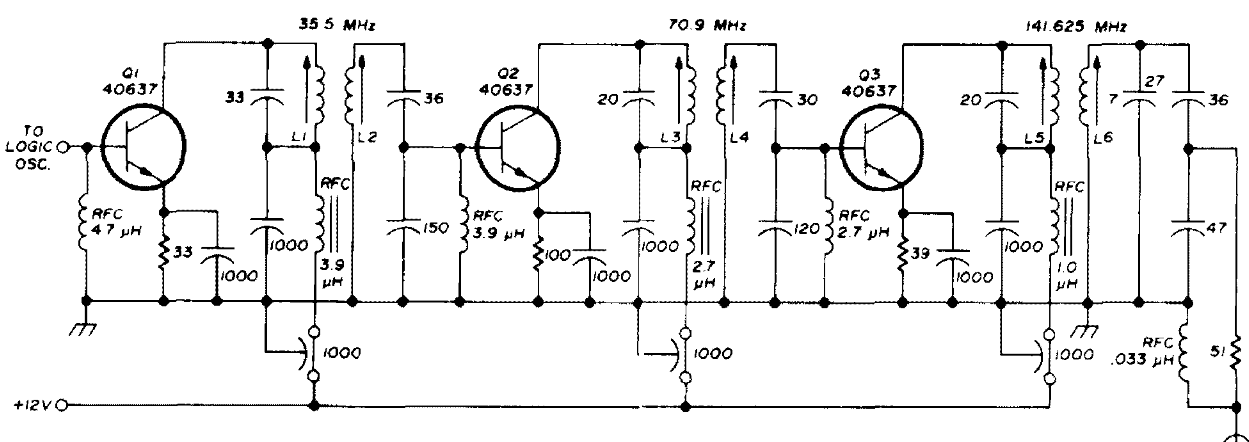


L1,L2	6 turns number 22, air wound, ¼" diameter
L3	6 turns number 22, air wound, ¼" diameter, center tapped
L4	37 turns number 32 on Amidon T50-2 toroid core

fig. 2. Simple two-meter converter for the two-band fm receiver. The ferrite bead on the drain lead of Q1 is an Amidon 45-101.

The frequency-selector switch is a 2-pole, 6-position rotary wafer switch wired so that +12 volts is applied to the two-meter converter when the two-meter channel crystals are switched into the circuit. More crystal frequencies can be added simply by adding additional logic oscillator stages — the only limiting factor to the number of logic-oscillator channels is the current handling ability of the voltage regulator.

The logic oscillator will operate properly with fundamental-mode crystals up to about 20 MHz. Above 20 MHz it is



L1,L2	11 turns number 22 on ¼" diameter slug-tuned form
L3,L4	4½ turns number 22 on ¼" diameter slug-tuned form
L5,L6	1½ turns number 20 on ¼" diameter slug-tuned form

fig. 4. Local oscillator multiplier chain. Stagger-tuned circuits provide relatively flat output across the two-meter fm band. L1, L3 and L5 are peaked for the lowest frequency crystal; L2, L4 and L6 are peaked for the highest.

necessary to use higher speed gates than the TTL ICs shown in **fig. 3.** Do not use overtone crystals in this circuit as they will not oscillate at the same frequency as that marked on the crystal can.

spaced by the diameter of one coil form (¼inch). The input coil of each pair is peaked for the lowest frequency crystal while the secondary coils are peaked for the highest frequency crystal. The fre-

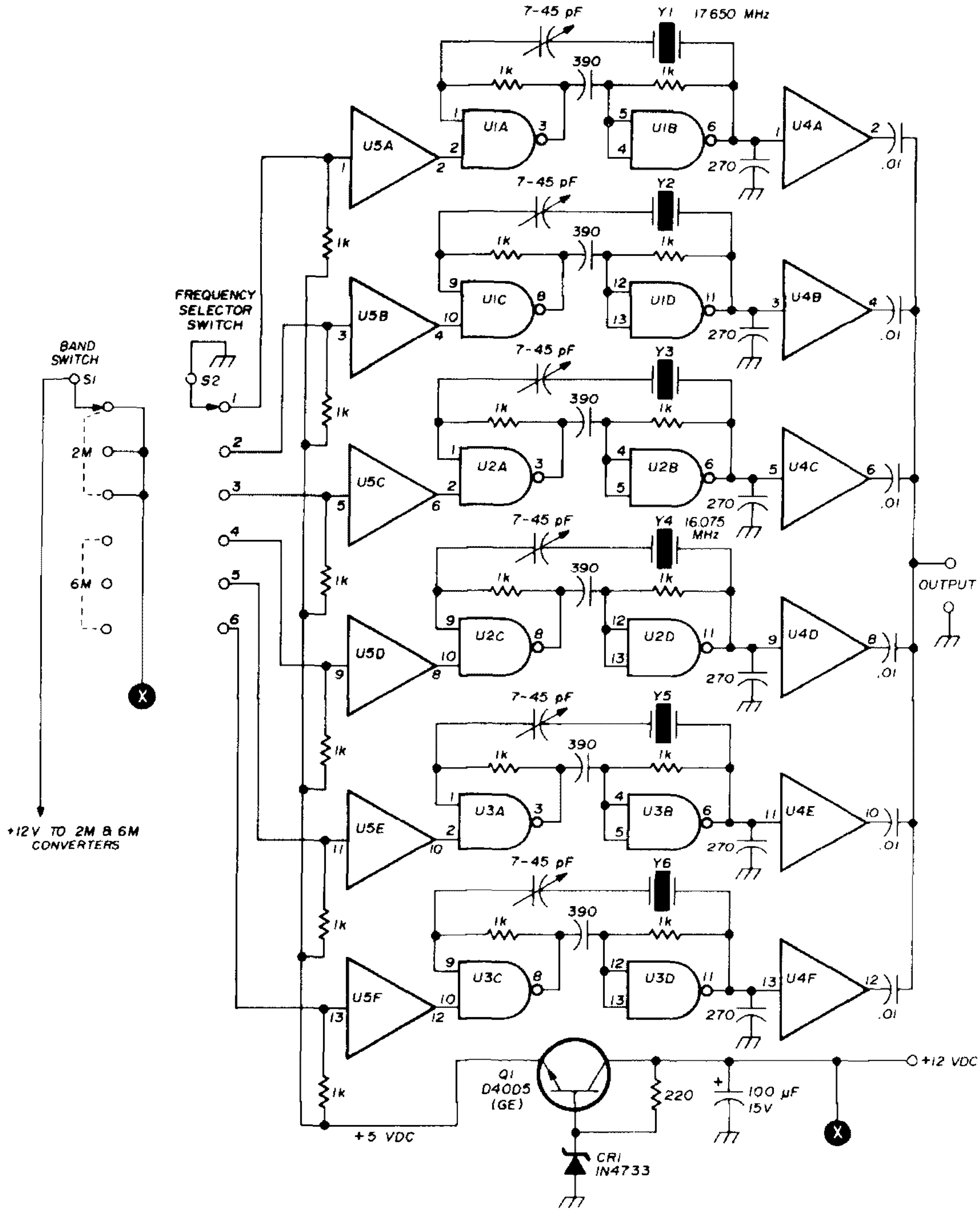


fig. 3. The logic oscillator circuit uses TTL ICs which are suitable for use with fundamental crystals up to approximately 20 MHz. Overtone crystals do not operate properly in this circuit.

Construction of the local-oscillator chain (**fig. 4**) is very straight forward and should cause no problems. The stagger-tuned stages provide the bandwidth necessary to cover the entire two-meter fm band. Each of the inductors is wound as described in **fig. 4,** and the coil pairs quencies indicated in the circuit diagram are the approximate center frequencies of the stagger-tuned stages.

The two-meter converter, logic oscillator and multiplier chain are built on a single piece of copper-clad board 3-inches wide by 8-inches long. The oscillator is

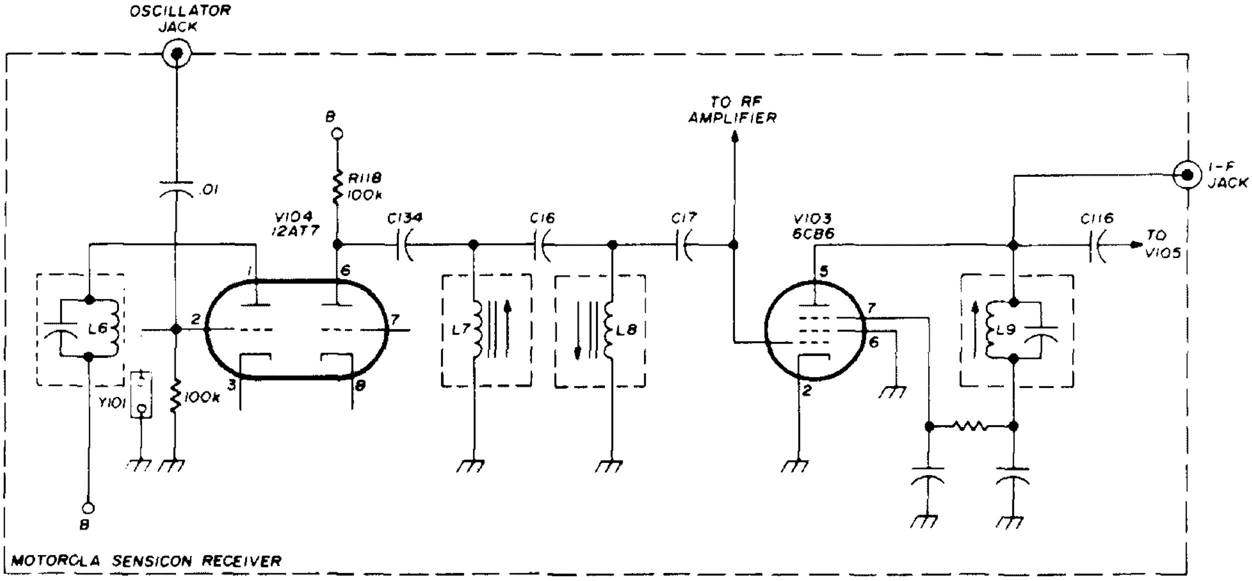


fig. 5. Partial diagram of the Motorola Sensicon receiver strip used in the two-band fm receiver built by W1SNN.

built into a separate shielded compartment as are the multiplier chain and the two-meter rf amplifier and mixer stages. Short lengths of coaxial cable are used to connect these units together and to the fm receiver strip.

receiver strip

The low-band Motorola Sensicon receiver (model PA9244-12) I modified for use in the two-band receiver is easily moved into the six meter fm band by replacing the fixed tuning capacitors in the rf amplifier, mixer and local-oscillator stages. The values for these capacitors are given in the Motorola schematic and are prefixed with the letters L, M or H, depending on the desired operating frequency. The H values (for high-band) are the values that should be installed for six-meters.

To use the receiver strip in the two-band fm receiver it is necessary to add an i-f input jack and an access jack to the first local oscillator. The i-f jack is located close to V103 so that the lead to pin 5 of the 6CB6 is as short as possible (see **fig. 5**). Another jack is mounted on the opposite side of the chassis and connected to pin 2 of V104, the grid of the 12AT7 oscillator multiplier. This completes the modifications to the receiver strip.

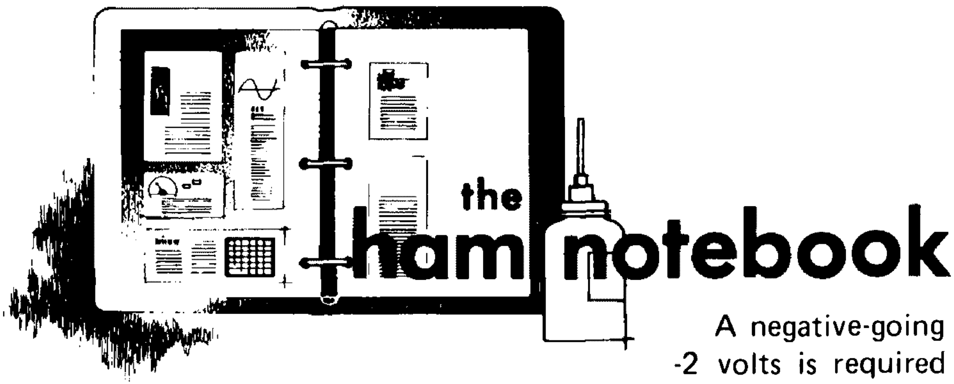
The receiver is aligned by connecting a center-scale dc vtvm to the discriminator output and adjusting the frequency control trimmer of each channel crystal until the meter reads zero with an incoming signal.

references

1. Stirling Olberg, W1SNN, "Logic Oscillator for Multi-Channel Crystal Control on VHF FM," *ham radio*, June, 1973, page 46.
2. Sherman M. Wolf, "FM Schematic Digest," P.O. Box 535, Lexingtin, Massachusetts 02173.

ham radio

"Delta . . . India . . . Sierra . . . Hotel . . . Echo . . . Sierra."



vhf fm scanner modifications

Although solid-state equipment is fast becoming the ideal equipment, at the same time there is still quite a bit of tube-type equipment in use. At the time I saw the February, 1973, edition of *ham radio* I was in the process of trying to come up with a scanner to add to my base station receiver. After looking over K2ZLG's vhf fm receiver scanner, I decided it was worth a try. However, the basic design was not completely acceptable for use with tube-type equipment. The following modifications were developed and tried. So far the unit has worked flawlessly.

Since most vacuum-tube receivers produce a negative-going voltage when the squelch is open, the original input circuit will not work. The circuit shown in fig. 1 was finally tried and seemed to work the best of any. One of the major problems with the bipolar input was to get the input impedance high enough to prevent loading of the receiver squelch circuitry; the dual-gate fet takes care of this problem nicely.

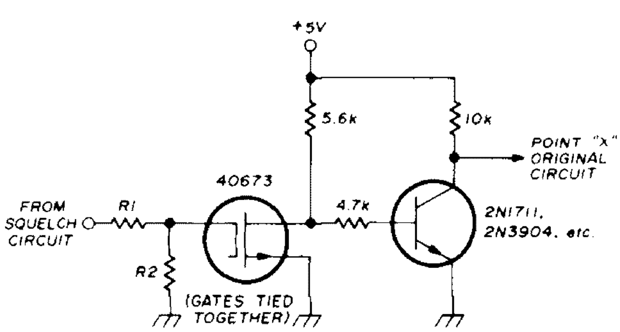


fig. 1. Input circuit for frequency scanner.

A negative-going voltage of more than -2 volts is required to stop the scanner. However, this voltage should not be more than -6 volts at the gate of the fet. For voltages in the range of -2 to -6 volts R2 can be eliminated. For voltages higher than -6 volts, R1 and R2 should limit the gate voltage to less than -6 volts. Typical values are between 1 and 10 megohms. The important thing is to try to keep the input impedance as high as possible since

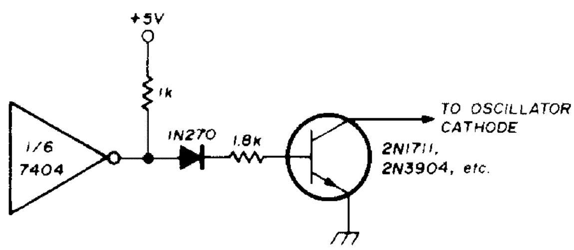


fig. 2. Output circuit. Repeat for each channel.

loading of less than about 1 megohm will interfere with normal squelch operation.

The next problem is that most tube-type receivers use separate oscillators for each frequency. Most circuits require that the cathode of the appropriate oscillator be grounded for operation of that particular oscillator. In addition, the ungrounded cathode produces about 30 volts. The simplest way around this seems to be to use an npn transistor to isolate the cathode from the scanner since TTL circuitry will not normally tolerate 30 volts. This is shown in fig. 2. The transistor used is non-critical as long as it will withstand the voltage present when the tube cathode is above ground.

In the unit at my station some MC4039 ICs were on hand instead of the 7446 decoder, so these were used. This IC happens to have an enable pin that was used with an inverting transistor off the

output of the input circuit. I didn't like seeing the numbers go by as the unit scanned, and this configuration turns the readout off while the unit is scanning. The same thing could be done by putting a switching transistor in the +5-volt line to the readout.

Mike Jones, WA5WOU

10 MHz coverage for the SB-303

The utility of 15-MHz WWV reception on Heath SB-303 receivers is somewhat dubious, considering present propagation conditions in this area. I've changed mine to tune WWV-10, finding virtual 24-hour coverage on 10 MHz. Modifications are relatively simple, but refer to the manual and schematic.

The 23.895 MHz crystal, Y104, used for 15 MHz, is replaced with an 18.895 MHz, HC6-U type, third overtone crystal intended for a 32 pF load. This change is made on the *crystal switch-board* (85-348), whose Xray pictorial with other PC boards is found at the back of the SB-303 manual.

To resonate the LC circuit marked "15 MHz" on the *heterodyne oscillator switch-board*, a 33-pF dipped mica or disc capacitor is added across C131. The slug in L117 must be moved in a few turns until oscillation occurs as noted by voltage appearing at TP on the PC board. An extra half-turn provides positive crystal starting when switching bands.

Modification of the *rf amplifier switch-board* (85-346) involves isolating foil pad areas around switch-points 5 and 6. Switch-point 5 will then be jumpered back to the foil lead coming from L111, the 14-MHz tuned circuit. A new resonant circuit for 10 MHz is required. I used 22 turns of number-24 enamel wire on an Amidon T-50-2 toroid, turns spaced evenly around the core. This is approximately 2.9 μH which, with a 47-pF disc in parallel, will resonate – out of the circuit – at 14 MHz. The older vacuum tube type of grid dipper will dip this unit satisfactorily. One end of this LC combination is soldered to switch-point 6 (on 85-346) and the other to ground foil near the *rf in* phone jack. Mount it close to the board, avoiding shorts.

Operation of the receiver on 10 MHz may be checked by attaching a short antenna through a few pF to *rf in* at C106 on the *amplifier switch-board*. The *preselector* should resonate broadly at about 30 to 40 percent of its range.

The *antenna switch-board* (85-345) is modified similarly to the previous PC board. Again, switch-point 5 and 6 pad areas are isolated, 5 being jumpered back to the foil lead running to the 14-MHz tuned circuit L103-C103. Also, between and clear of switch-points 6 and 7, drill about a number-58 hole. This board has a double-section rotary switch, the section nearest the board being the secondary, and the outer, the primary. Switch-points 5 and 6 on the primary are isolated by unsoldering the blue wire and jumper between 5 and 6. Resolder the blue wire directly to switch-point 5 (14 MHz).

Primary switch-point 6 is left blank for the moment. The LC antenna circuit also uses an Amidon T-50-2 toroid with 22 turns of number-24, with the addition of 6 turns of number-26 or -28 wire forming the primary. Use an adjacent winding rather than over-winding for the primary. The tuning capacitor is again a 47-pF disc paralleled with the 22-turn secondary. The combination mounts on the foil side of the board (85-345).

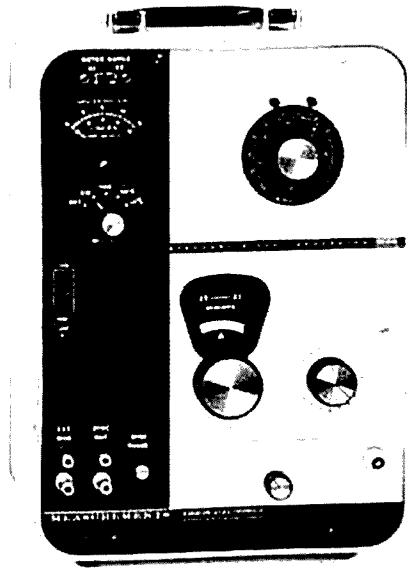
One end of the secondary goes to switch-point 6 and the other to any convenient ground-foil point. One side of the 6-turn primary also ties to this point, the other side being fed through the pre-drilled hole and soldered to outer switch-point 6. An adjustment of a turn on the secondary may be desirable for better tracking, but I found the *pre-selector tuning* to be adequately sharp.

This application can be used for other 500-kHz segments. The crystal frequency must be 8895 kHz above the lowest signal frequency.

Bill Fishback, W1JE

new products

fm signal generator



The Measurements Model 800A series of solid-state fm signal generators cover all mobile communication frequency bands allocated by the FCC. Any desired frequency can be quickly obtained by selecting one of the six frequency bands, tuning the coarse tuning control, and making narrowband adjustments with either the electronic fine tuning or incremental frequency controls.

The Model 800A signal generator provides accurate output voltages traceable to the National Bureau of Standards. The output is continuously variable from 0.1 microvolt to 0.1 volt by means of a mutual inductance type attenuator. Output voltages are automatically maintained at all levels by a temperature-compensated bolometer circuit. Rf leakage is negligible, and microphonics are so low that accurate receiver sensitivity measurements can be made down to 0.1 microvolt.

Internal modulators provide frequency modulation at 1000 Hz sine wave or 20 Hz sawtooth. External modulation from dc to 30 kHz may be applied through front-panel binding posts. Sync out and sync phase are available for external modulation (up to ±32 kHz peak deviation) so that dual-trace sweep alignment techniques may be used.

For complete technical data write to Edison Electronics, Division of McGraw-Edison Co., Grenier Field, Manchester, New Hampshire 03103, or use *check-off* on page 94.

WWV data folder

Complete, up-to-date information on the many services provided by The National Bureau of Standards Radio Stations WWV, WWVH and WWVB is being offered at no charge by the True Time Instrument Company, manufacturers of receivers for all of the standard time and frequency broadcasts.

The NBS transmissions provide an invaluable service to radio amateurs, laboratories and engineers throughout the world. Extremely precise audio and radio frequency standards are broadcast, as well as accurate time signals, geophysical alerts, Atlantic and Pacific area storm warnings and radio frequency propagation forecasts. This information is at the disposal of anyone having a receiver capable of tuning to one or more of the transmitting frequencies. The proper use of NBS facilities can greatly supplement the instrumentation of any laboratory.

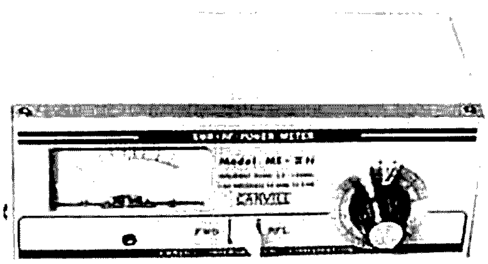
Maximum utilization of this valuable "natural resource" depends upon a complete knowledge of the current broadcasting schedules and transmitting frequencies. The folder supplies this data, as

More Details? CHECK–OFF Page 94

well as information on suitable methods of comparison with local chronometers or instrumentation. Also included are the hourly broadcast schedules of National Bureau of Standards stations WWV, WWVH, WWVB, with supporting data.

Write to True Time Instrument Company, 225 Melbrook Way, Santa Rosa, California 95405, and request Bulletin 373-1, or use *check-off* on page 94.

swr meter



Carvill International has announced its new in-line swr and power meter, the model ME-IIN. The ME-IIN is a direct-reading swr and power meter which measures the ratio of the forward and reflected wave on a coaxial transmission line. In this instrument a printed-circuit transmission line is used to eliminate unbalanced rf pickup which is often a problem in more simple swr meters. The swr meter is usable on all bands from 3.5 to 150 MHz.

For more information, write to Carvill International Corporation, 825 Constitution Drive, Foster City, California 94404, or use *check-off* on page 94.

signal intensifier

The new SABA-5 (Symtek Automatic Broadband Amplifier) provides low-noise and high *useful* gain for amateur communications receivers. This new amplifier, which covers 80, 40, 20, 15 and 10 meters with no tuning has a typical noise figure of 2.5 dB and gain of 20 dB (minimum). Input and output impedance is 50 ohms.

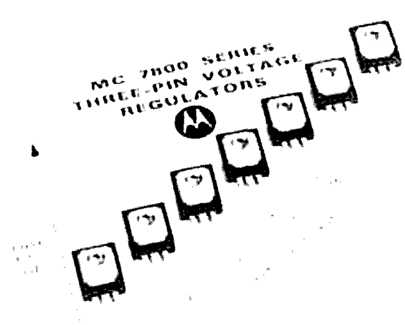
The SABA-5 uses a dual-gate, diode-protected mosfet to take advantage of its low noise characteristics as well as its

inherent protection against strong rf fields. The Qs of the tuned circuits are engineered for high gain and adequate bandwidth to cover the entire amateur bands with MARS overlap. Insertion loss, when the amplifier is turned off, is kept to a minimum through the use of a straight-through coaxial loop and low-loss ceramic switches.

Biasing of the amplifier is optimized for best gain, cross-modulation and noise-figure characteristics. Gain may be adjusted by varying the main supply voltage ±3 volts. The amplifier is easily installed on any communications receiver in minutes by simply inserting it between the receiver and the antenna. Transceivers may be easily modified by breaking the antenna circuit from the T/R relay to the receive rf amplifier and bringing each end to an external phono jack.

The SABA-5 carries a 30-day money-back guarantee and 1-year warranty, and is priced at $79.95. Models for 160 meters, 6 meters and 2 meters are also available. For more information, write to Symtek, Inc., Box 128, Clearwater, Florida 33517, or use *check-off* on page 94.

three-pin voltage regulators



Many times the need arises for a simple, low-cost voltage regulator which can provide a moderate amount of current without complex current-boosting circuitry. Applications include on-card regulation and power supply distribution in large systems.

A new Motorola device family composed of seven fixed-voltage regulators

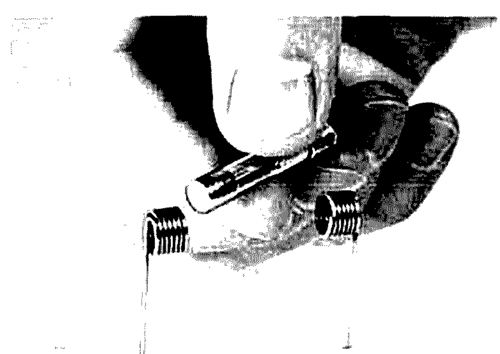
More Details? CHECK—OFF Page 94

housed in a popular plastic power transistor package fulfills these needs. The MC7805/24 series positive voltage regulators can supply in excess of 1 ampere at nominal voltages of 5, 6, 8, 12, 15, 18 or 24 volts (as designated by the last two digits of the device number). However, unlike most voltage-regulator ICs, these devices have only three terminals – input, output and ground – and they require no external components. The devices can be easily attached to a heatsink surface with a machine screw through the hole in the package to attain higher maximum power dissipation.

To insure a rugged device, internal current limiting, thermal shutdown, and output transistor safe operating area compensation techniques are employed. These features make the regulators essentially burn-out safe.

For further information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, Post Office Box 20912, Phoenix, Arizona 85036, or use *check-off* on page 94.

spring-type fuse holder



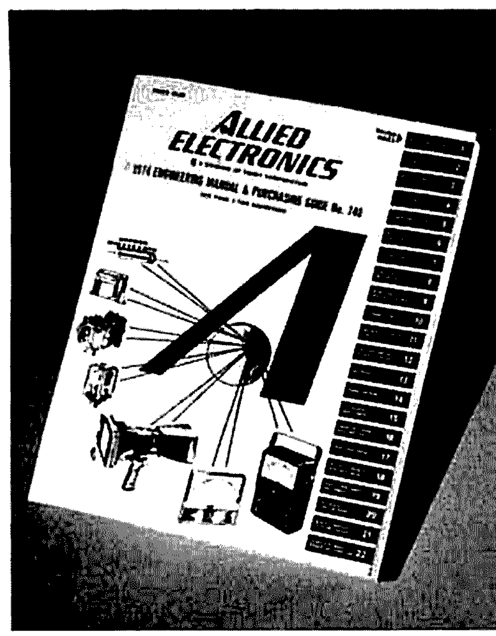
Oneida Electronics has recently introduced a new coil-spring fuse holder that makes it easier than ever to replace fuses. The new holder eliminates the need for using more costly type pig-tail fuses and does away with cutting and re-soldering pig-tail leads.

Regular fuses can be snapped into the coil-spring holder in seconds. Service people will find them ideal for use in those hard-to-get at places. Replaces permanently installed pig-tail fuses by merely soldering the leads of the new

spring holder to stubs of removed pig-tail fuse. The spring holders have been designed to accommodate tv, radio, hi-fi and most other electronic device fuses.

The new spring-type holder is permanent — quality constructed of tempered spring steel with dip soldered leads. Available packaged 5 pair per pack on dealer cords and in bulk for OEM use. For detailed information write Oneida Electronic Manufacturing, Inc., Meadville, Pennsylvania 16335, or use *check-off* on page 94.

new allied catalog



Allied Electronics (Division of Tandy Corporation) has published their new catalog number 740. Previous catalogs have served as the electronics industry's "answer book," and the new catalog is even better. In addition to the easy-to-use tab-index format and easy-to-use 9 x 11-inch size introduced in 1973, even more useful product information is included in the book. Prime feature of the Engineering Manual and Purchasing Guide catalog is the inclusion of *Engineering Drawings* of all electrical components. All physical dimensions are given to allow efficient design of electronic packages before components are purchased. Electrical characteristics of all items are also included.

Allied has also introduced a new policy for obtaining a copy of their

More Details? CHECK—OFF Page 94

catalog: instead of the $5.00 price, or $10.00 order requirement, *anyone* can now obtain a copy for the cost of postage and handling – just $1.00. All items shown are in stock at all Allied warehouses. With Allied enjoying the best order filling record in the industry, this, as always, is the one catalog you can't do without. For your copy, send $1.00 for postage and handling to Allied Electronics, 2400 W. Washington Boulevard, Chicago, Illinois 60612.

transistor substitution handbook

The *Transistor Substitution Handbook* is updated continuously, and a new edition is published annually. This 13th edition has been published in an easy-to-read 8-1/2x11-inch format, and contains over 100,000 transistor substitutions. To guarantee the most accurate possible substitutions, the electrical and physical parameters as described in the manufacturers' published specifications for each bipolar transistor were fed into a computer; then each transistor was compared with all others. Consequently, transistors which matched within prescribed limits are listed as substitutes.

Section 1 of the handbook contains substitutions for both American and foreign-made transistors which are arranged in numerical and alphabetical order. Types recommended by the manufacturers of general-purpose replacement transistors are included at the end of each list of substitutes. Additional data on these general-purpose replacement types – manufacturer, npn or pnp, germanium or silicon, and the recommended applications – are also reviewed.

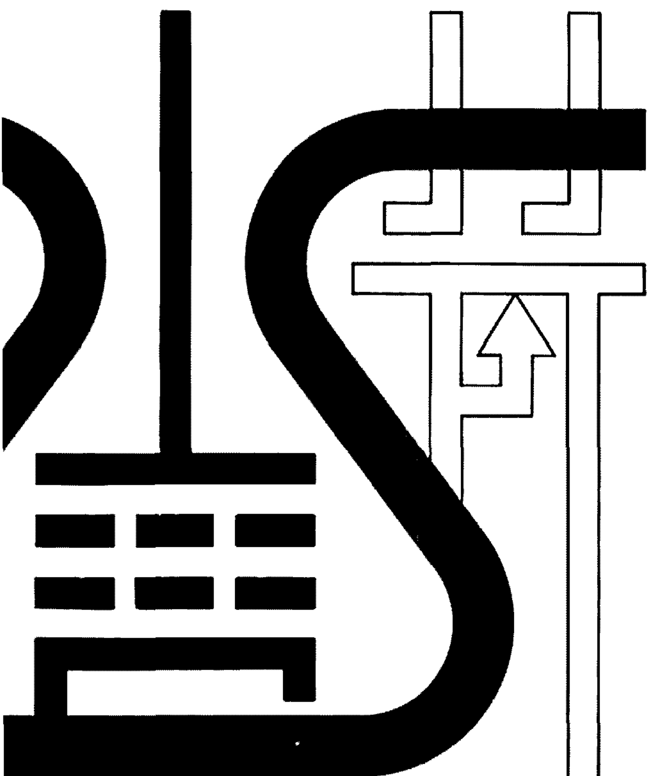
The *Transistor Substitution Handbook* is a valuable source of information for amateurs concerned with transistor replacement in communications industrial, commercial or home-entertainment equipment. 144 pages, softbound. $2.95 from Comtec Books, Greenville, New Hampshire 03048.

85 cents

focus on communications technology . . .

ham radio magazine

MARCH 1974



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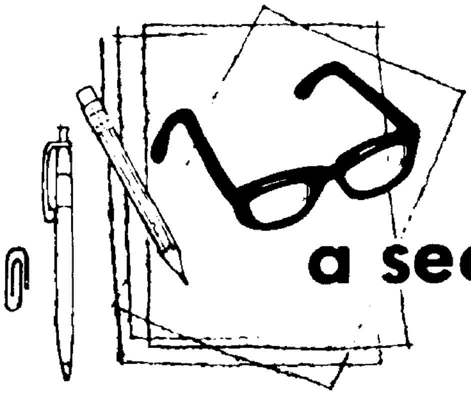


a second look

by jim fisk

A few months ago, down in Oklahoma, several old-time radio amateurs, each now retired, attended an auction of the electronics equipment and *junk* collection of a prominent local amateur. From all reports, it was quite a collection, filling four large warehouses. Except for the huge volume (and original cost) the collection resembled the typical "hell box" of every amateur who lived through the halcyon days when building your own transmitter was conventional practice and everyone eagerly added to his junk collection at every possible opportunity. The same is still true to a somewhat lesser extent, with amateurs maintaining large collections of old and new electronic goodies for some, yet unplanned, project. Talk to any amateur who has been around for a few years and you're sure to find a garage, an attic or a basement full!

However, there was one big difference in the Oklahoma collection. Where most radio amateurs painfully part with dollars, this amateur had painlessly parted with thousands of dollars. The contents of the four warehouses vividly reflected this difference. Think back a few years – what was the most delectable piece of radio gear you could imagine? It was probably in the Oklahoma collection. And not just one, but several. Parts, radio sets, test equipment, you name it, it was all there in unimaginable profusion. One whole warehouse floor was crammed full of big transmitters, spark coils and rotary gaps for 1920-style transmitters, spiderweb coils and thousands of variable capacitors of every possible make and description. The list could go on for pages.

Now here's the tragedy: These priceless articles, which belong in a museum, were grouped in huge lots with utter junk and sold to the only people who can handle large lots of junk – junk dealers! The probability that these dealers have the background to differentiate between the valuable antique and the valueless junk is frighteningly small. Antiques that can never be replaced, items not preserved in any collection, are going to be bulldozed under at a county landfill dump, and that's a bloody shame.

This scene, on a much more modest scale, is probably repeated many times a year. Without getting morbid, each one of us should realize that we are not immortal. Each of us has a collection of electronic gear that we've acquired over the years that will, if someone doesn't know any better, be bulldozed under with the trash at the city dump when we join the list of Silent Keys. Each item, when you acquired it, represented a jewel to be treasured and was carefully put away. If you were ever so careless as to toss out one of these treasures, you could be sure you would have an almost immediate pressing need for an identical article. I know, because it's happened to me everytime I've cleaned house!

The point is this: Talk to your heirs. Clue them in as to what items belong in a museum. Better yet, make arrangements with the executor of your estate to donate certain prized items to the museum of your choice. This applies not only to equipment, but to your library of old electronics books. The same sort of foresight applies to your newer equipment as well. Give your executor the names of several trusted amateur friends who will help dispose of modern radio gear and test equipment. They will know the fair market value – the executor may not. There have been more than a few cases where an amateur's survivors have been ripped off to the tune of thousands of dollars. Don't let it happen to your family.

Jim Fisk, W1DTY
editor-in-chief

simple ssb transmitter and receiver for 40 meters

Complete construction information for a no-frills ssb and CW transceiver system that offers high performance at low cost

W.J. Weiser, M.D., VE3GSD, 98 Banstock Drive, Willowdale, Ontario, Canada

This article describes a simple, high-performance 40-meter ssb and CW receiver and transmitter. The receiver incorporates a very stable vfo and incremental tuning while the transmitter features 180-watt PEP input. Although the receiver was designed to operate in conjunction with a companion 40-meter exciter, it can also be used with any number of separate exciters.

Furthermore, the basic design can be used on any of the amateur bands from 80 through 10 meters by simply using the appropriate tuned circuits in the rf and mixer stages, and by changing the frequency of the vfo so it tunes 455 kHz below the desired band. For maximum stability vfo pre-mixing should be considered for the two higher-frequency bands, 10 and 15 meters.

the receiver

A look at the block diagram in **fig. 1** shows that the basic receiver circuit is a rather conventional single-conversion super-heterodyne. The incoming 7-MHz signal is coupled to the first rf amplifier, Q1, amplified, and applied to gate 1 of the mixer, Q2 (**fig. 2**). The dual-gate mosfets used in the rf amplifier and mixer stages provide high gain, low noise and a minimum of cross-modulation problems. These particular devices are also internally gate protected, a significant bonus when the receiver is used in strong rf fields. Diodes CR1 and CR2 are included in the input circuit as additional protection to the first rf stage.

The 6.545- to 6.845-MHz vfo signal is injected at gate 2 of the mixer. The 455-kHz difference frequency is selected by the i-f transformer, T1. The very stable vfo circuit features incremental tuning and output buffering, and is a modification of an earlier design (**fig. 3**).[1] Output buffering is provided by the emitter-follower stages Q6 and Q7. Transistor Q7 also serves as a power amplifier, allowing the vfo to drive a 7360 balanced mixer tube used in the companion exciter.

The incremental tuning circuit uses a Motorola MV1654 varactor diode, CR3. This circuit permits an approximately 10-kHz offset to either side of the vfo frequency. The control voltage on the varactor is set by R3, the offset tuning control. Resistor R4 compensates for differences in varactors and allows the vfo to be adjusted to zero offset.

Switch S2 allows the receive offset to be activated manually; receiver offset is automatically turned off by relay K1 when the receiver is placed in the transmit or standby mode. This particular type of switching is necessary because a 1.5-kHz frequency difference between the transmitted and received signal exists at my station. This frequency difference resulted from the use of a 455-kHz mechanical filter in the exciter and a 453.5 kHz filter in the receiver. If you are building from scratch, I recommend that described receiver (fig. 6).[2] The use of a CA3028A as the product detector allows significant conversion gain in this stage. The high-impedance output of the CA3028A is matched to the base of the audio preamplifier transistor, Q8, through transformer T5. Although the Motorola MC1454 provides more than sufficient audio output, an MC1554 or HEP593 may be substituted for even more audio. Alternately, a Motorola MFC9010 can be used to replace both Q8 and U2 and deliver about 2-watts of audio.[3]

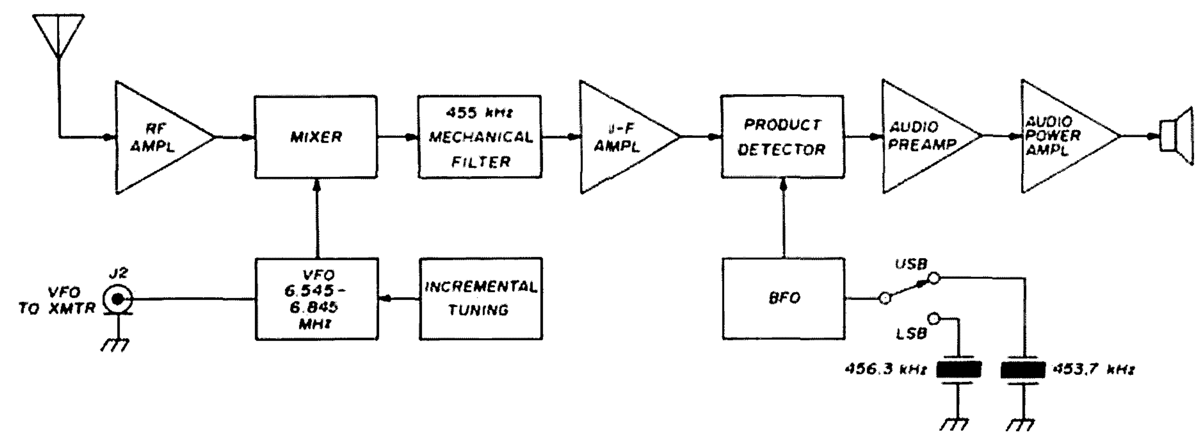


fig. 1. Block diagram of the single-conversion 40-meter receiver ssb and CW receiver.

you use a 455-kHz filter in the receiver, such as the Collins FA21-7102.

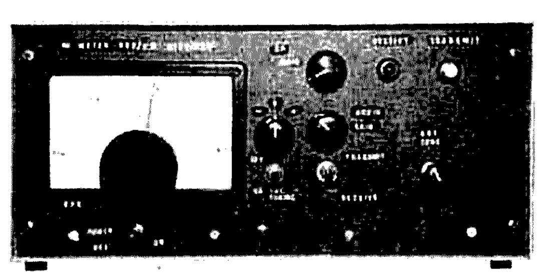
The two-stage i-f amplifier, Q3 and Q4, is relatively straightforward and provides more than enough i-f gain (fig. 4). For the sake of simplicity I did not include an agc system in the receiver – if you want to include agc, it should be connected to gate 2 of the rf and i-f amplifier stages, Q1, Q3 and Q4.

An MPF102 fet is used in a simple bfo circuit which is based on the old tuned-grid, tuned-plate vacuum-tube circuit (fig. 5). Bfo output is taken from the secondary of transformer T4. For optimum performance of the CA3028A product detector, U1, the amplitude of the bfo injection voltage should be 2 to 3 volts rms. The sideband crystals are selected by switch S1. The crystal frequencies shown in the diagram are for use with a Collins 455-kHz mechanical filter.

The product detector and audio stages used here were adapted from a recently

In my receiver with the audio gain control, R2, adjusted for maximum gain, the MC1454 was driven into oscillation by transistor Q8. This was corrected by bypassing a small amount of the input audio signal to ground through C7 (0.1 μF). Because of component differences, and differences in circuit layout, you may not require this bypass capacitor.

Any well-regulated power supply with an output of 12 to 13.5 volts may be used with this receiver. Excellent voltage



Homebrew 40-meter receiver uses all solid-state circuits.

regulation is required to ensure maximum vfo stability because the MV1654 varactor used in the incremental tuning system will significantly shift the vfo output frequency with the slightest dc voltage variation. For additional vfo stability it might be a good idea to insert a three-terminal voltage regulator IC, such as the Fairchild μA7812, in the vfo supply line.[4]

receiver construction

The mechanical layout of the receiver is shown in the photographs. The rf amplifier, mixer and mechanical filter are constructed on a 2¼x2¼-inch section of Vector board, copper clad on one side. The i-f amplifier is built on another, similar sized board, and the vfo and incremental tuning are built on another 2¼-inch square board. These three boards are mounted on 3/4-inch spacers under the chassis.

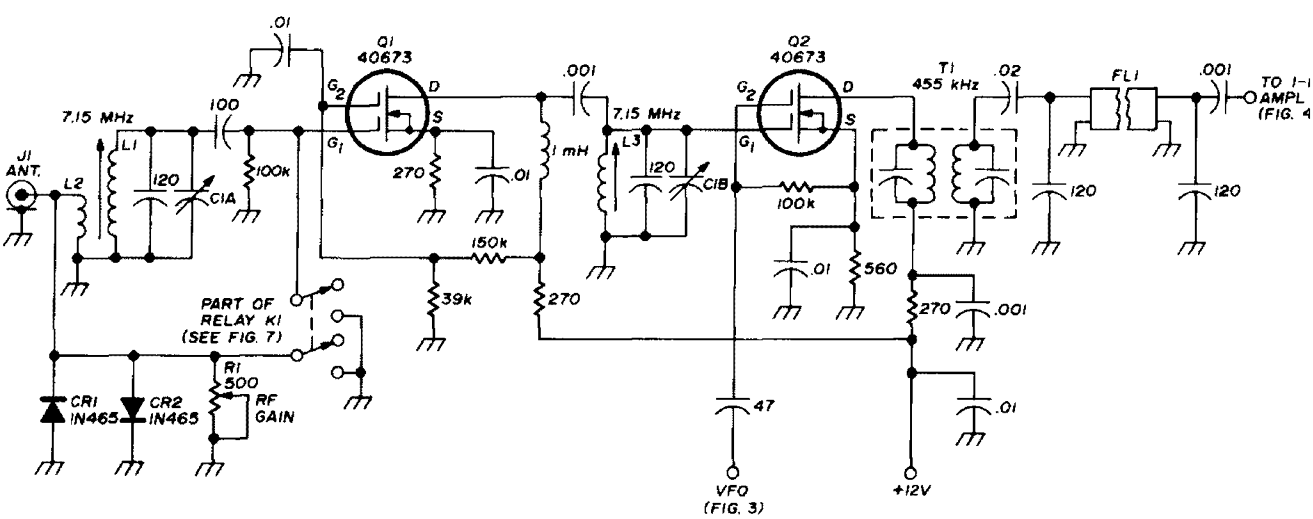
The bfo, product detector, and audio stages are built on a 2½x5-inch board which is mounted on top of the chassis on 1-3/4-inch spacers. Relay K1 is installed under this board. Each board is built and tested individually, and connected to the others with miniature shielded cable.

The 1-inch-deep aluminum chassis, 10-inches long and 5½-inches deep, is installed in a commercial enclosure. The rf amplifier and mixer tuning capacitor, C1, is a modified dual-section 365-pF broadcast variable with all but two rotor plates removed from each section. With the values given for C8 and C9, the modified variable will cover the frequency range from 7.0 to 7.3 MHz.

receiver alignment

Before starting the alignment procedure, all slug-tuned coils must be roughly adjusted to the proper frequencies with the help of a grid-dip meter. Also, each stage is initially tested and aligned before being mounted on the chassis. This points up any difficulties that might be more difficult to pinpoint when the receiver is completely assembled. Inductor L4 in the vfo is adjusted for an output of 6.695 MHz with C4 set at mid-excursion.

Although a signal generator is best for initial alignment of the rf and mixer stages, it is possible to use an on-the-air signal. Apply a 7.15-MHz signal to L1 and peak C1 for maximum rf voltage on gate 1 of the rf amplifier stage. Next apply a 455-kHz signal to the primary of T1 and tune the transformer for maximum signal



C1	modified dual-section 365-pF broad-cast variable (see text)	L1,L3	30 turns no. 26 wound on ¼" slug-tuned ceramic form, resonated to 7.15 MHz
FL1	455-kHz mechanical filter, 2.1-kHz bandpass (Collins FA21-7102)	L2	4 turns no. 26 on cold end of L1
		T1	455-kHz input i-f transformer

fig. 2. Schematic diagram of the rf amplifier, mixer and mechanical filter.

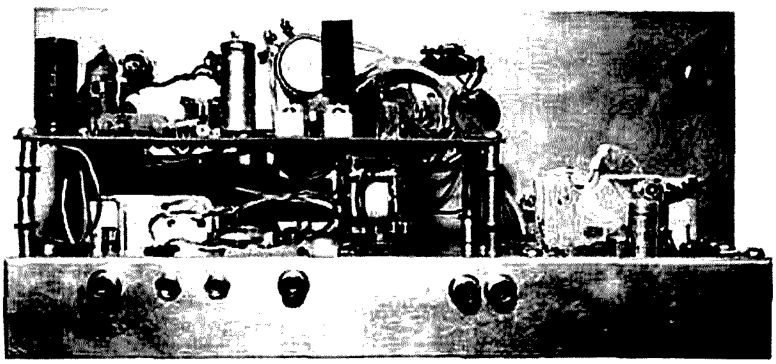
at the output of the mechanical filter. Since a reading of 0.1 volt is typical here, a sensitive rf probe and voltmeter is required.

To align the i-f amplifier a 455-kHz signal is applied to gate 1 of transistor Q3 and transformers T2 and T3 are tuned for maximum signal at the output of T3. The bfo should require no adjustment although T4 may be tuned, as required, if the bfo fails to oscillate.

After each of the individual boards has been tested and aligned, and the receiver has been assembled, final peaking can be accomplished with an on-the-air signal. The rf amplifier, mixer and i-f stages are tuned for maximum receiver gain. To obtain good tracking with C1 it is necessary to alternately re-tune the mixer and rf amplifier coils, L1 and L3, several times. The vfo inductor, L4, is accurately adjusted and the tuning dial calibrated with the aid of a communications receiver equipped with a crystal calibrator.

To align the incremental tuning system, monitor the received frequency, adjust resistor R3 and note the amount of frequency change. A 40% change in the resistance of R3 should result in an approximate 20-kHz shift in the vfo frequency. The dc voltage range coinciding with this frequency shift is measured at the wiper of R3 and should be in the range from zero to about 2.5 volts.

When the dc voltage required for a 20-kHz frequency shift has been determined, the wiper of resistor R4 is set to the midpoint of this dc voltage swing. With the incremental tuning turned off, vfo output is centered at its mid-frequency point with R4, and resistor R3 will move the vfo about 10 kHz to either side of this center point.



Rear view of the 40-meter receiver. The product detector and audio board is mounted on 1-3/4" spacers. Relay K1 is underneath the board while the vfo tuning capacitor is on the right.

receiver performance

The performance of the receiver is most rewarding. It appears to be as sensitive as my FTDX-401, and its frequency stability is excellent. No significant vfo warmup drift has been noted, the audio is apparently distortion free, and the set provides more than enough gain. One very pleasing feature of the mosfet stages is their very low noise figure. The ambient noise level in this receiver is the lowest of any comparable receiver I have ever used.

the transmitter

The 180-watt PEP ssb transmitter may be used as a separate unit, or with the receiver. The unit is completely self-contained and incorporates a stable vfo, power supply and all the necessary control functions for antenna switching and receiver muting. The use of an RCA 7360 beam-deflection tube for the balanced modulator and balanced mixer stages provides excellent carrier suppression and local-oscillator rejection, two requirements that are difficult to achieve in single-conversion ssb exciters which use a 455-kHz i-f.

A block diagram of the transmitter is

shown in fig. 8. The output of the high-impedance microphone is amplified by V101, and applied to one of the deflection electrodes of V102, the 7360 balanced modulator. The crystal-controlled carrier signal is injected at the cathode (see fig. 9). The 455-kHz grid 1 and grid 2) serves as a self-oscillatory carrier generator, with switch S101A selecting either Y101, a 455-kHz crystal for tuneup or CW, or Y102, 456.25-kHz crystal for lower-sideband operation. For tuneup or CW one of the deflection plates of V102 is grounded by

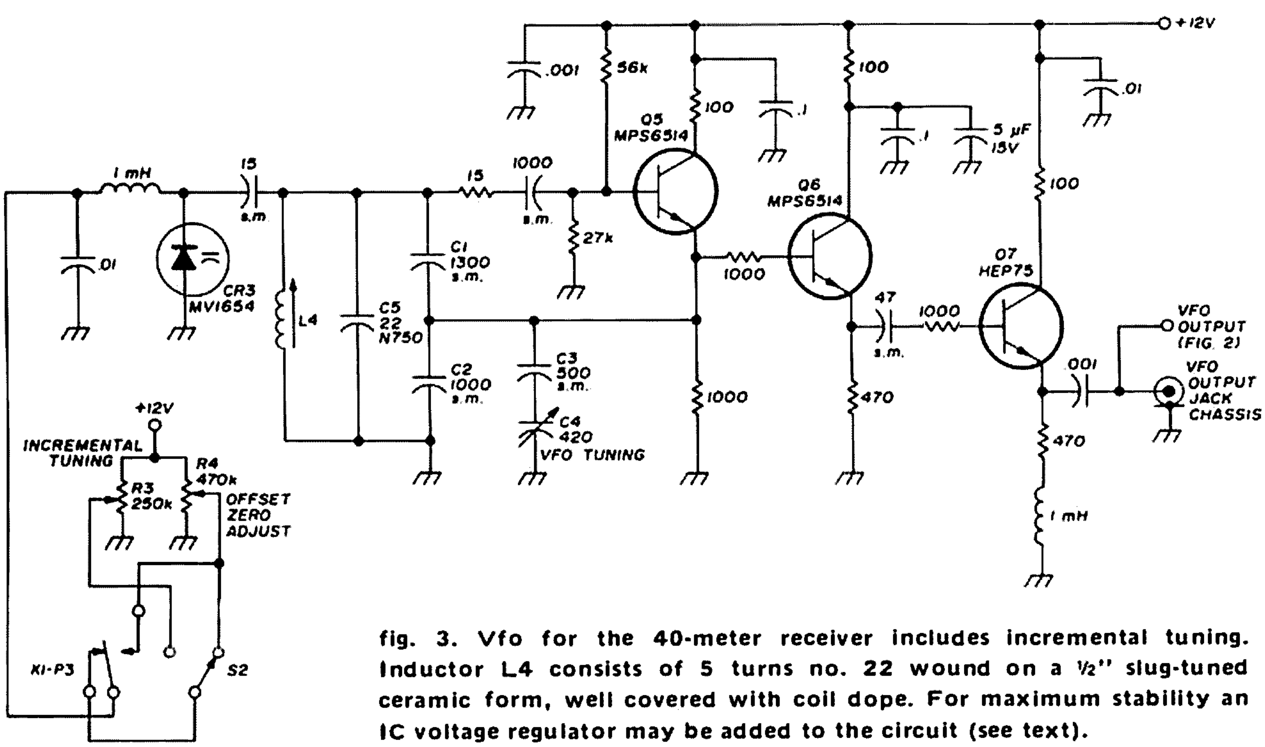
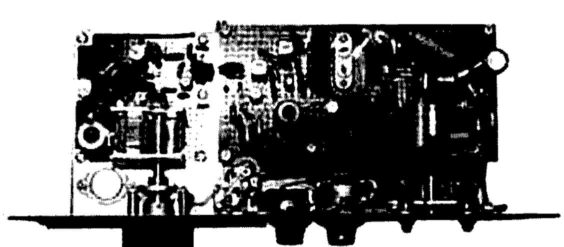


fig. 3. Vfo for the 40-meter receiver includes incremental tuning. Inductor L4 consists of 5 turns no. 22 wound on a ½" slug-tuned ceramic form, well covered with coil dope. For maximum stability an IC voltage regulator may be added to the circuit (see text).

double-sideband output signal is coupled into a Collins mechanical filter.

One of the absolute requirements of a modern single-sideband exciter is adequate carrier suppression. By using a 7360 beam-deflection tube up to 60-dB of carrier suppression can be achieved if careful construction and layout have been followed, and the circuit has been properly adjusted.

The triode section of V120 (cathode,



Top view of the receiver chassis. The vfo board is to the left, the product detector and audio board to the right.

switch S101B, effectively unbalancing the modulator so that a 455-kHz carrier signal appears across the output network. In ssb service the audio signal applied to one of the deflection plates unbalances the modulator.

Another requirement for effective and courteous single-sideband operation is adequate suppression of the unwanted sideband. The Collins mechanical filter can provide up to 60-dB of sideband attenuation if careful attention is given to mechanical layout, and stray coupling is prevented between the input and output of the filter. The filter also attenuates the 455-kHz carrier signal by an additional 20 dB. The 120-pF capacitor across the output of the filter helps match the output impedance of the filter to the input impedance of the 6BA6 i-f amplifier, V103 (see fig. 10).

The ssb signal from the output of the 6BA6 i-f amplifier is coupled through

transformer T101 to V104, the 7360 balanced mixer. The triode section of V104 is used in a Colpitts-type internal vfo. Capacitor C102 is the main tuning capacitor and C103 and C104 serve both as frequency-trimming and temperature-compensating capacitors. For transceive short as possible and the input and output must be sufficiently isolated; the 33-ohm non-inductive resistor in series with the grid serves as a parasitic suppressor.

The screen bypass capacitor is soldered directly across the base of the 6GK6

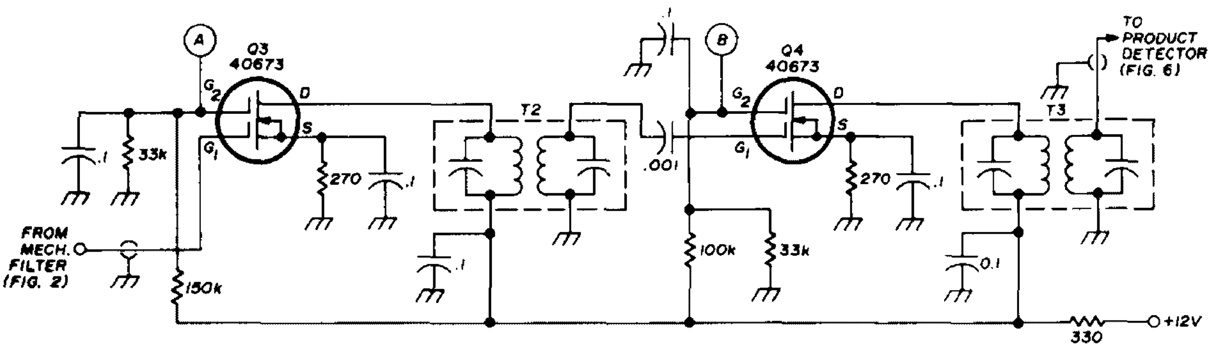


fig. 4. Two-stage 455-kHz i-f amplifier. Transformers T2 and T3 are miniature 455-kHz input i-f transformers.

operation the internal vfo is bypassed and an external vfo coupled into grid 1 through switch S102.

One problem when designing a 40-meter ssb exciter using a 6.545-MHz vfo and a 455-kHz i-f in a single-conversion system is inadequate vfo suppression. Since the vfo frequency is only separated from the desired output frequency by 455 kHz, a single-tuned resonant circuit in the output of a conventional mixer stage would probably not be inadequate to sufficiently attenuate the vfo signal.* However, by using a 7360 beam-deflection tube as a balanced mixer, up to 40-dB of vfo rejection is possible.

The 7.0- to 7.3-MHz ssb output from V104 is coupled through L102-L103 to the 6GK6 driver stage, V105, **fig. 11.** A 6GK6 was selected as the driver because it offers considerable gain and can safely handle a 300-volt plate supply. However, because of its high gain, care must be taken to prevent instability and self-oscillation. All leads must be kept as socket between the input and output pins and the ground end is also soldered to the central pin of the socket. This technique provides an effective grounded shield between the input and output circuits of the driver stage.

The output of the 6GK6 driver is coupled into the grids of V106 and V107, the 6146B power-amplifier tubes through C106 and L104. Capacitor C106 should be pruned by removing plates from a miniature variable capacitor until the tuned circuit resonates from 7.0 to 7.3 MHz with one full revolution of the shaft. Fix-tuning the power amplifier's grid circuit to the desired band makes tuneup simple and prevents the operator from inadvertently tuning the final amplifier to some unwanted spurious signal that may appear at the output of the high-gain

*Both single- and double-tuned bandpass circuits were tried by the author. The double-tuned bandpass arrangement was workable with a conventional 6BA7 mixer stage, but it allowed an unacceptably high level of vfo signal to feedthrough to the driver stage.

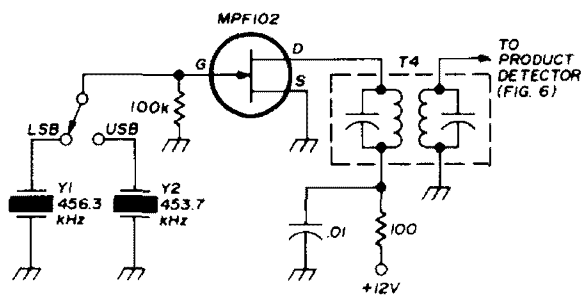


fig. 5. The 455-kHz bfo stage. T4 is a miniature 455-kHz i-f input transformer.

driver stage. Resistor R105, a 5000-ohm, 5-watt wirewound potentiometer, serves as the driver gain control.

A pair of 6146Bs in parallel are used as a power amplifier (**fig. 12**). These tubes are still among the best power amplifier tubes for ssb service, offering good linearity, excellent IMD characteristics and good power sensitivity. They can also withstand considerable abuse during groggy-eyed, early morning DX chasing tune ups.

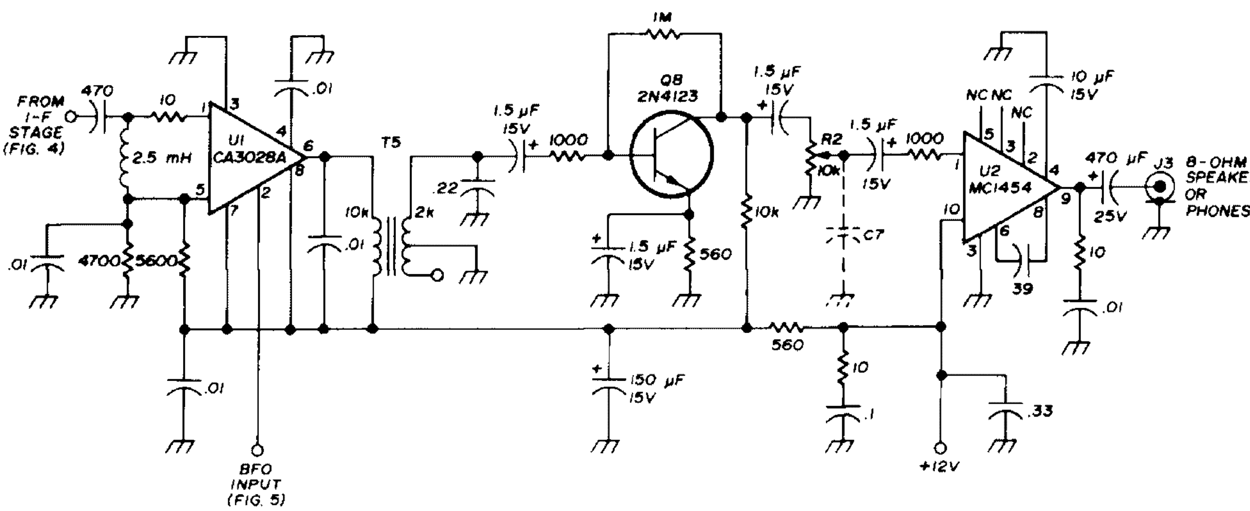


fig. 6. Product detector, audio preamplifier and audio output stage. Capacitor C7 (0.1 µF) may be required to suppress an oscillating output stage (see text).

As with the driver stage, considerable care must be taken in keeping leads short around the PA stage sockets, separating input and output circuits, and shielding the cathode keying line to prevent self-oscillation. Parasitic suppressors Z101 and Z102 help minimize any high-frequency instability. With these techniques, I did not find it necessary to neutralize either the driver or PA stages. Both are extremely stable.

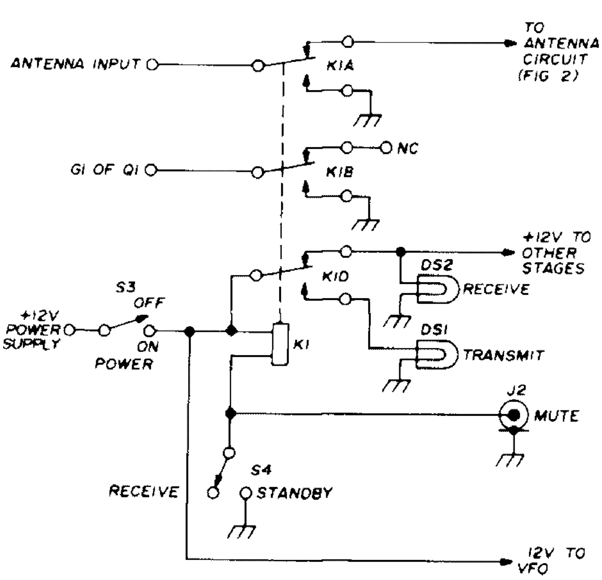


fig. 7. Switching and control circuits for the 40-meter receiver. Relay K1 is a miniature 4-pole relay with a 12-volt coil.

The final-amplifier tank circuit is a standard pi network. Capacitor C108 is a surplus, high-voltage 400-pF variable. With the values specified for C107 and L105, the stage loads nicely into a 50-ohm resistive load. You may wish to change the value of L105. In that case an additional capacitance (C109, 100-400 pF) may be needed in parallel with C108 to load properly. Or, if you wish, a triple-section 365-pF broadcast variable, with all sections in parallel, may be substituted for C108.

In the interest of simplicity and thriftiness, a grid and plate current metering system was excluded in favor of a simple rf output indicator. A sample of the output rf is applied to a 1N34A diode through a voltage-dividing network; the rectified dc is filtered and read on a 0-1 milliameter. Resistor R106 serves as a sensitivity control.

transmitter control circuits

Since this exciter was intended to work in conjunction with the matching receiver, antenna switching and receiver muting functions were included in the

design. All control functions are provided by two surplus, 6800-ohm, 55 Vdc relays. One relay, K101, switches antennas and functions in receiver muting while the other, K102, is used for B+ switching between transmit and standby (see fig. 13).

ternal vfo, to talk himself on frequency using the companion receiver as a monitor.

power supply

The 350-0-350-volt center-tapped winding of transformer T102 is used in a

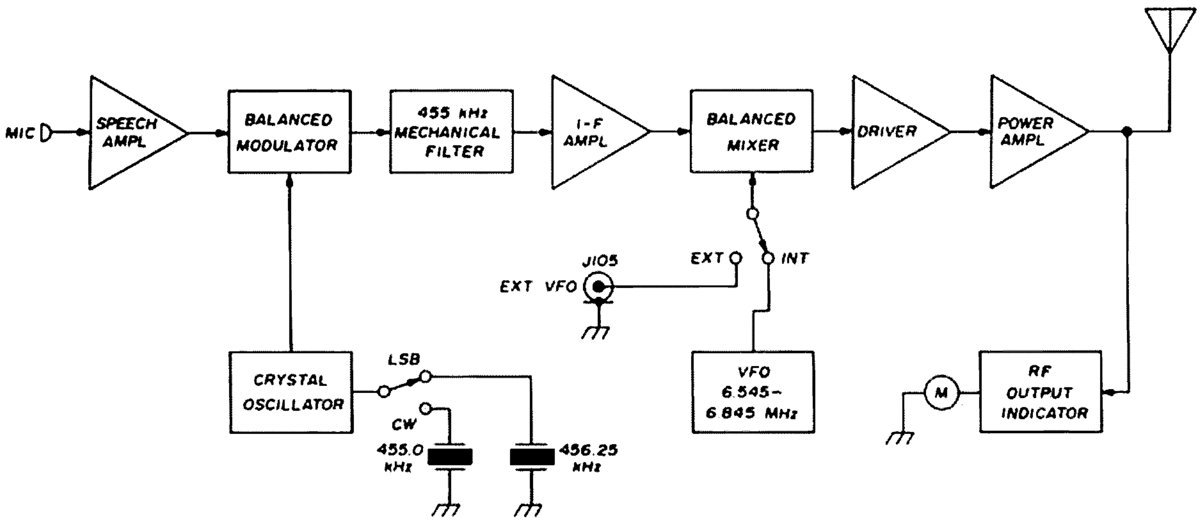


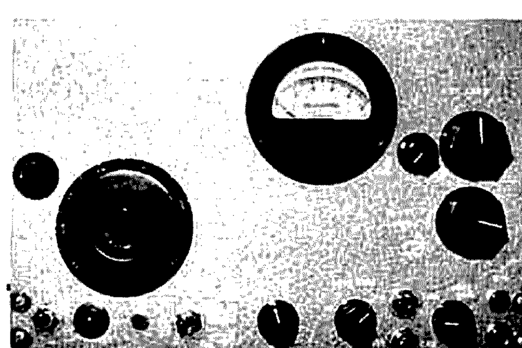
fig. 8. Block diagram of the 40-meter ssb and CW transmitter. Power input is about 180-watts PEP.

One relay section, K101A, switches the antenna between the receiver and exciter while K101B grounds the exciter output during standby. A third section, K101C, activates the receiver's muting relay during transmission.

The second relay switches the 210- and 310-volt B+ supplies to the final-amplifier tubes and the driver, respectively, via K102C and K102A. Another section, K102B, switches 210-volts to the speech amplifier, V101, and i-f amplifier, V103, during transmit. This B+ switching is paralleled by a second manual switching network, S104, which is used as a spotting switch.

With S104 in the *transmit* position, activating the PTT switch through S101 or the manual transmit/tune switch, S103 picks up the relays and switches the exciter from standby to transmit. With S104 in the *spot* position the PTT switch and S103 are interrupted and no relay switching can occur, but B+ is applied to the speech amplifier and i-f amplifier through S104B. This spotting function allows the operator, when using the in-

conventional full-wave bridge circuit to provide 750-volts dc for the final amplifier (see fig. 14). The 310-volt supply for the driver is obtained from the center tap after choke-input filtering. The regulated 210-volt supply is provided by a pair of OB2 regulators in series. Bias voltage for the 6146Bs was acquired by reversing a 6.3-volt filament transformer and rectifying the 117-volt winding. A 50k, 10-watt wirewound potentiometer, R5, is used as the bias voltage adjust control (see fig. 14).



Front panel of the 40-meter ssb and CW transmitter. Power input is approximately 180 watts.

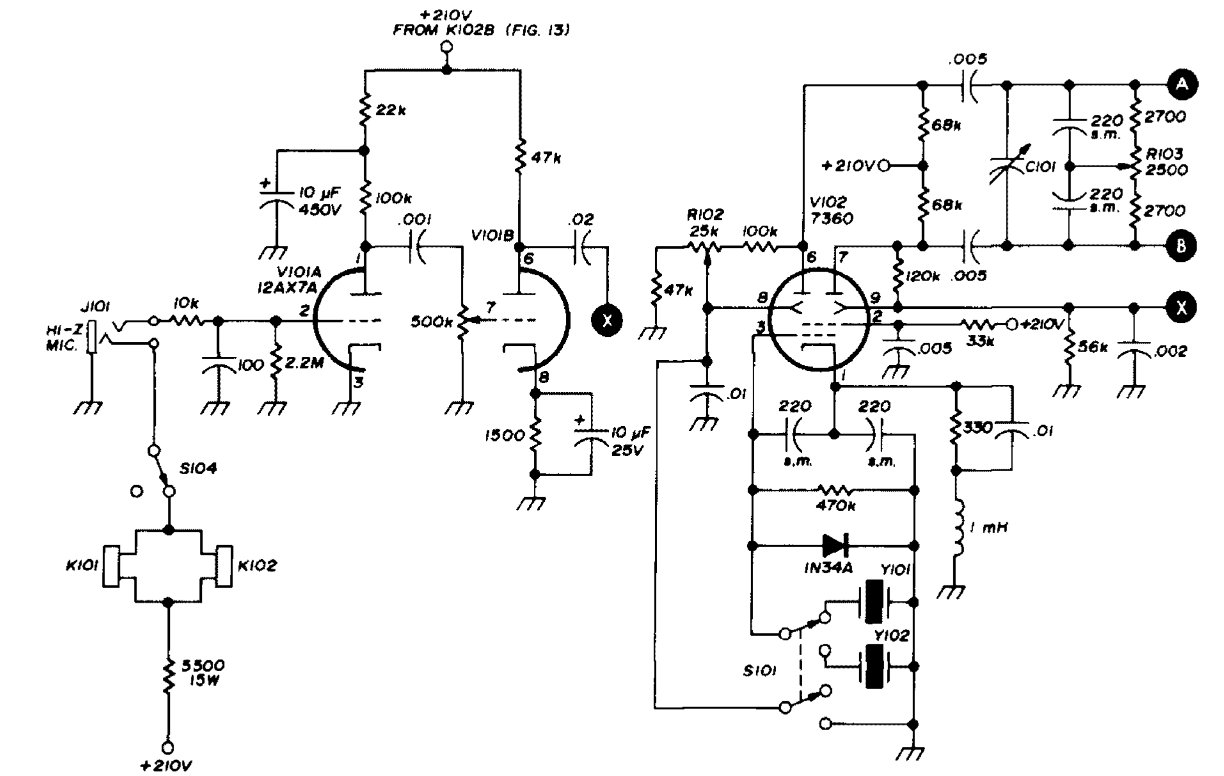
transmitter construction

The eternal frustration of any ardent home-constructor today is finding the needed parts. Most of the values given in this design can be varied; all the tuned-circuit inductors and capacitors of the tuned circuits, with the exception of the vfo, can be changed, of course, if required resonant frequencies are maintained. The seasoned builder may have his own scheme of chassis layout and wiring, and neither is particularly critical.

My transmitter was built into a 12x10x2-inch aluminum chassis. The photographs show the layout. In any chassis layout a straight-line approach is usually best and that is what I used, with the speech amplifier followed by the balanced modulator, the mechanical filter, the i-f amplifier and so on.

The power amplifier cage entirely encloses the final-amplifier tubes and the pi network and is made of 1/16-inch-thick aluminum siding with a ventilated top cover. The plate-tuning and antenna-loading capacitor shafts are brought out to the front panel with 1/4-inch couplers and shaft-extenders. For the sake of neatness and compactness I used a rather elaborate system of universal joints and gears to angle the shaft of capacitor C107 away from the speech amplifier tube, V101. This system can be avoided if you use a larger chassis or disrupt the straight-line layout by moving V101's socket closer to the edge of the chassis.

Two variable capacitors, C106 and C102, were modified for use in this design. The grid-tuning capacitor, C106, was originally a 25-pF miniature variable; all but one rotor plate was removed to achieve the desired resonance range of 7.0



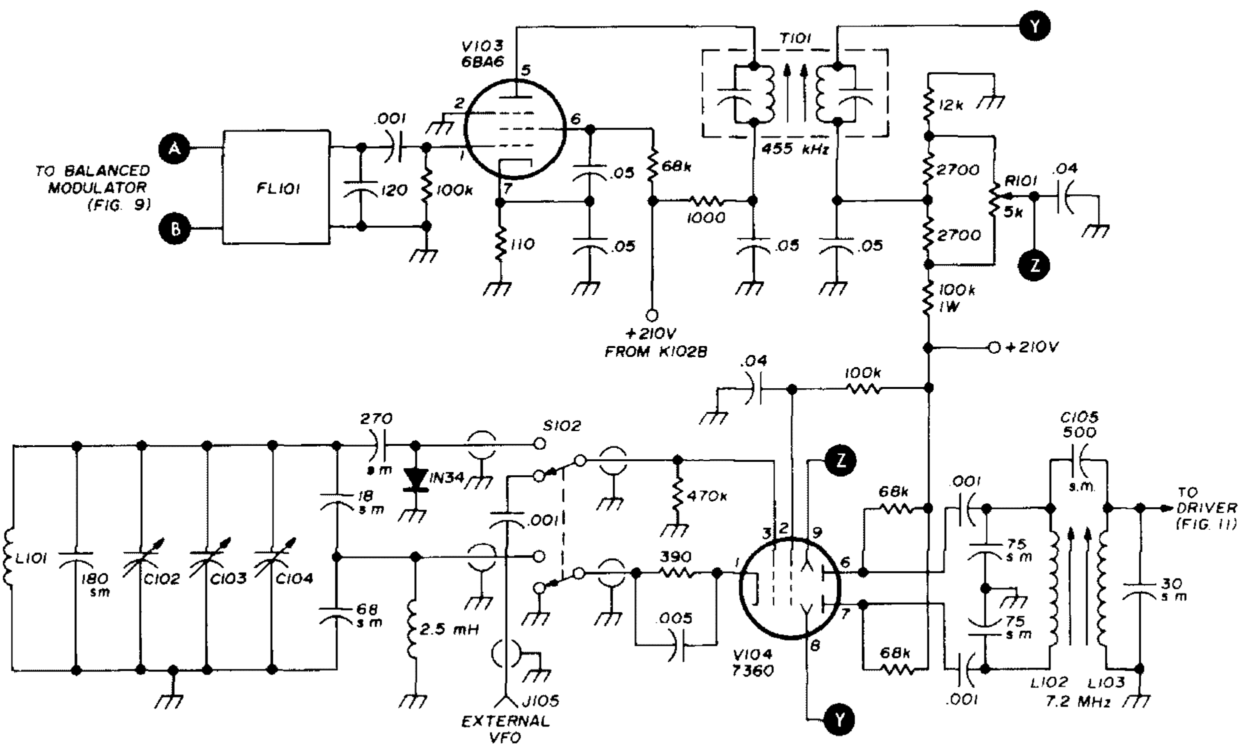
C101	5-25 pF NPO ceramic trimmer (phase balance control)	S101	dpdt toggle switch (lower-sideband, carrier)
K101,K102	surplus 4-pole relays, 55-volt, 6800-ohm coil	S104	dpdt toggle switch (spotting switch)
R102	25k potentiometer (carrier amplitude control)	Y101	455.0-kHz crystal (carrier)
R103	2500-ohm potentiometer	Y102	456.25-kHz crystal (lower sideband)

fig. 9. Speech amplifier and balanced modulator circuits. Complete relay switching system is shown in fig. 12.

to 7.3 MHz for one complete rotation. A small L-shaped bracket is used to mount C106 underneath the chassis.

The vfo main-tuning capacitor, C102, is a modified variable from a surplus ARC-5 receiver. The gear train and gear reduction ratio used on these capacitors make them ideal for amateur vfo applications. Only one capacitor section is used in the vfo. Each capacitor section had about 100-pF capacitance, and plates were removed until the entire excursion of C102 yielded a vfo range of 6.545 to 6.845 MHz with C103 and C104 set to midrange.

All the vfo components were securely soldered with short leads to firm tie points. Miniature shielded cable, soldered to frequent ground lugs, was liberally used in the switching circuitry around switch S102. The vfo tank coil was closewound on a 1/2-inch diameter ceramic form and Q-doped several times. Finally, the entire vfo was enclosed by an aluminum shield. The mechanical precautions were justified by the excellent vfo frequency stability characteristics I obtained.



C102	modified ARC-5 receiver tuning capacitor (see text)	L101	12 turns no. 22 enamelled on ½" diameter ceramic form
C103	4-25 pF NPO ceramic trimmer	L102,L103	29 turns no. 22 enamelled on ½" diameter, slug-tuned form
C104	4-25 pF N500 ceramic trimmer	S102	dpdt toggle switch (internal/external vfo)
C105	500-pF silver mica (see text for other values)	T101	455-kHz i-f input transformer
FL101	455-kHz mechanical filter (collins 455 FA21 7102)		

fig. 10. Transmitter vfo, i-f and balanced mixer circuits. Colpitts-type vfo circuit tunes from 6.545 to 6.845 MHz.

As mentioned previously, great care should be taken to isolate the input and output circuits of all rf stages. Particular attention should be paid to circuit isolation around the mechanical filter as stray capacitative coupling between the input and output circuits here can significantly reduce sideband and carrier suppression.

transmitter alignment

Aligning the exciter is simple and should be done in stages, with each stage

being built and aligned before the next one is started. The speech amplifier needs no adjustment and can be checked by connecting an ac voltmeter between the plate of V101 and ground through a series blocking capacitor. Speaking into a high-impedance microphone should produce a swing of several volts on the meter.

maximum carrier suppression. With a typical rf probe virtually no rf voltage should be recorded at the plate of V103 when the balanced modulator is properly adjusted.

Final adjustment of the i-f amplifier is now accomplished with switch S101 in the *carrier* position. Move the rf probe to the secondary of T101 and alternately

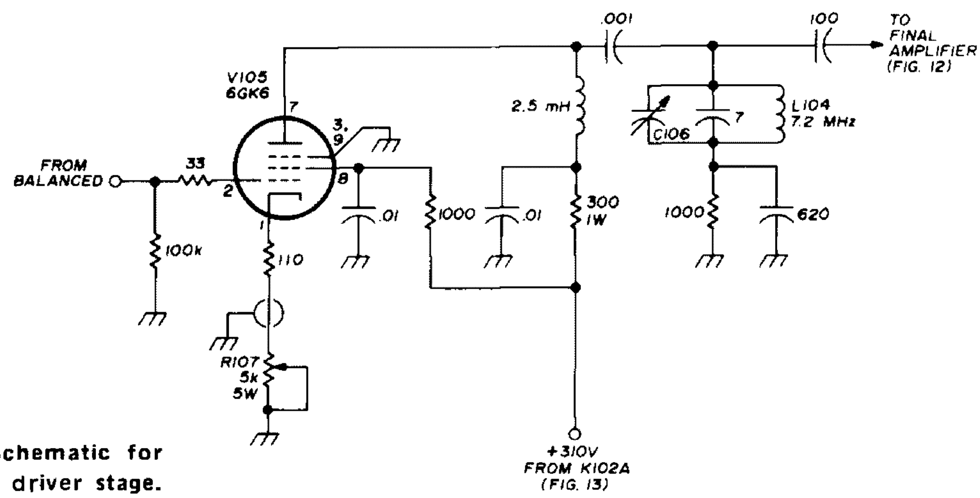


fig. 11. Schematic for the 6GK6 driver stage.

C106	approximately 5 pF (25-pF air variable with all plates removed except one
L104	28 turns no. 26 enamelled on ¼" slug-tuned form
R107	5000-ohm, 5-watt wirewound potentiometer

Since the mechanical filter has noticeable insertion loss, adjusting the balanced modulator is best done by monitoring the signal at the output of the 6BA6 i-f amplifier, V103. A vtvm with an rf probe is coupled to the plate of V103, and switch S101 is placed in the *carrier* position. This unbalances the modulator and allows the 455-kHz carrier oscillator signal to pass through the mechanical filter. About 5- to 10-volts rf should be recorded by the probe. Next, tune the primary of transformer T101 for maximum rf voltage.

To adjust the balanced modulator for optimum carrier suppression, switch S101 is switched to the *lower-sideband* position. This switches in the 456.25-kHz carrier oscillator crystal, Y102, and any rf voltage measured at the plate of V103 is due to carrier. Alternately adjusting R102 and C101, the amplitude and phase balance controls, for minimum rf voltage sets the balanced modulator, V102 for

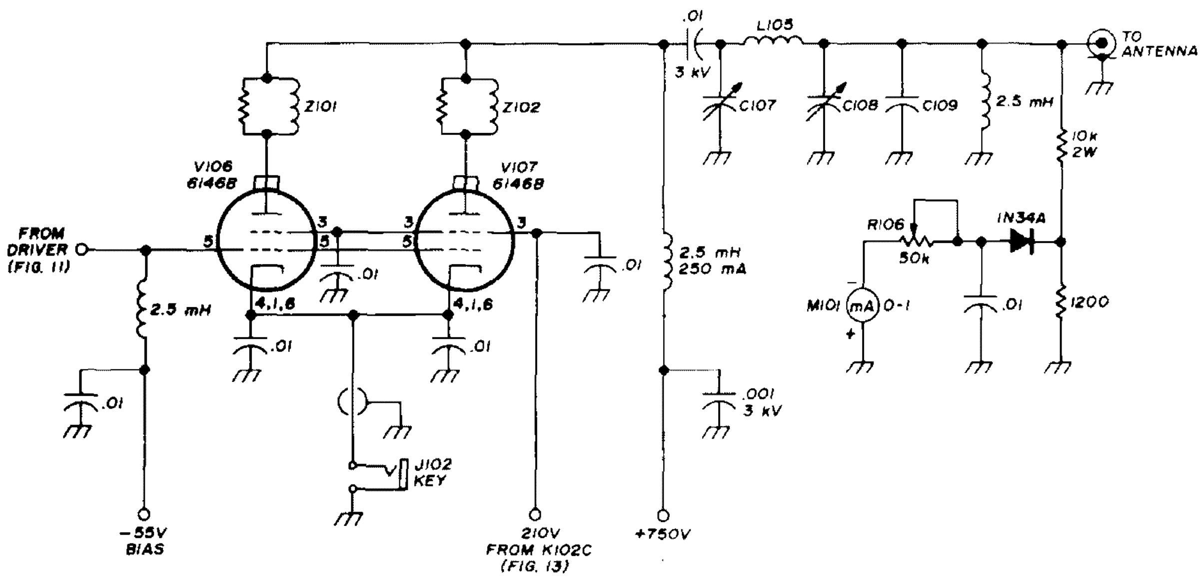
adjust the primary and secondaries of the i-f transformer for maximum rf voltage indication (typically 10 to 15 volts).

Putting the vfo on frequency can be most easily done by using a well-calibrated general-coverage receiver. If one is unavailable, a reasonably accurate grid-dip meter can be used to place the vfo within the proper general operating range. With capacitors C102, C103 and C104 set approximately at midexcursion, identify the vfo signal using a general-coverage receiver. Inductor L101 may have to be decreased or increased in value (by changing the number of turns) if the vfo signal is far from the desired range of 6.545 to 6.845 MHz. Now, with C102 at midrange, adjust C103 (and C104, if necessary) to bring the vfo frequency to 6.695 MHz. Temperature compensation was found to be unnecessary in my exciter as vfo warmup drift was quite acceptable and the frequency stability is excellent.

If you experience any excessive fre-

quency drift during warmup, this may be minimized by experimentally adjusting C104 to an increased value while checking on the frequency stability. Generally speaking, an uncompensated vfo will drift to a lower frequency as the temperature rises; adding a negative coefficient capacitance will minimize this. If it is necessary to increase the value of C104, be certain required, values of 10 to 100 pF can be used for C105, while still maintaining adequate mixer output.

Once the vfo has been set on the desired frequency, the balanced mixer can be easily aligned. During these adjustments, be certain that the internal vfo has been switched into the circuit. Place an rf probe on pin 2 of the empty driver socket



C107,C108	400-pF, 1000-volt air variable (see text)	M101	0-1 mA meter
C109	see text	R106	5000-ohm, 5-watt potentiometer (rf sensitivity control)
L105	23 turns no. 22 enamelled, closewound on 1" ceramic form	Z101,Z102	parasitic suppressors, 8 turns no. 16 enamelled, wound around 10-ohm, 1-watt carbon resistors

fig. 12. Power amplifier and rf output meter. Grid and plate current metering may be added if desired.

to reduce the value of C103 to keep the vfo within the proper operating range.

Tuned circuits L102 and L103 were not adjusted in the conventional bandpass manner as there was more than enough drive from the mixer when its output circuit was peaked at 7.2 MHz to fully power the 6GK6 driver across the entire 7.0- to 7.3-MHz range of the exciter. The coupling capacitor, C105, connected across L102 and L103 is 500 pF. The 7360 balanced mixer offered so much carrier oscillator rejection that I opted for maximum mixer output by heavily coupling L102 and L103 at the expense of decreased selectivity of these tuned circuits. However, if more vfo rejection is and, with S101 in the *tune* position and S104 in the *spot* position, alternately adjust L102 and L103 for maximum rf voltage. I measured more than 5 volts rf on the grid of V105 in my unit.

Maximum vfo carrier oscillator rejection is achieved in a similar fashion. Leaving the rf probe on pin 2 of V105, place S101 in the *lower-sideband* operating position, adjust R101, the rejection control, for minimum rf voltage. The 7360 balanced mixer is capable of about 40-dB carrier rejection and the measured voltage should be 0.05 volts or less with the proper adjustment of R101.

Before aligning the 6GK6 driver stage, be sure the final amplifier tubes, V106

and V107, are removed from their sockets. Place the rf probe on pin 5 of V106 (or V107) and set the driver control, R107, at midrange. Now place the exciter in the tune mode by switching S101 to the *tune* or *carrier* position and S104 to the *transmit* position. Move S103, the manual tune/transmit switch to the *tune* position, and with C106, the grid-tuning capacitor, at midrotation, adjust L104 for maximum rf voltage. Turning the driver output control, R107, should increase the

volts rf on voice peaks. The monitored, detected audio should be clear and crisp and without distortion.

Finally, switch the monitoring receiver to the upper sideband mode and check sideband suppression. No intelligible upper-sideband signal should be heard. With the sideband suppression check completed, all the low-level stages of the transmitter have been aligned and only the power amplifier and relative power output metering stages need be adjusted.

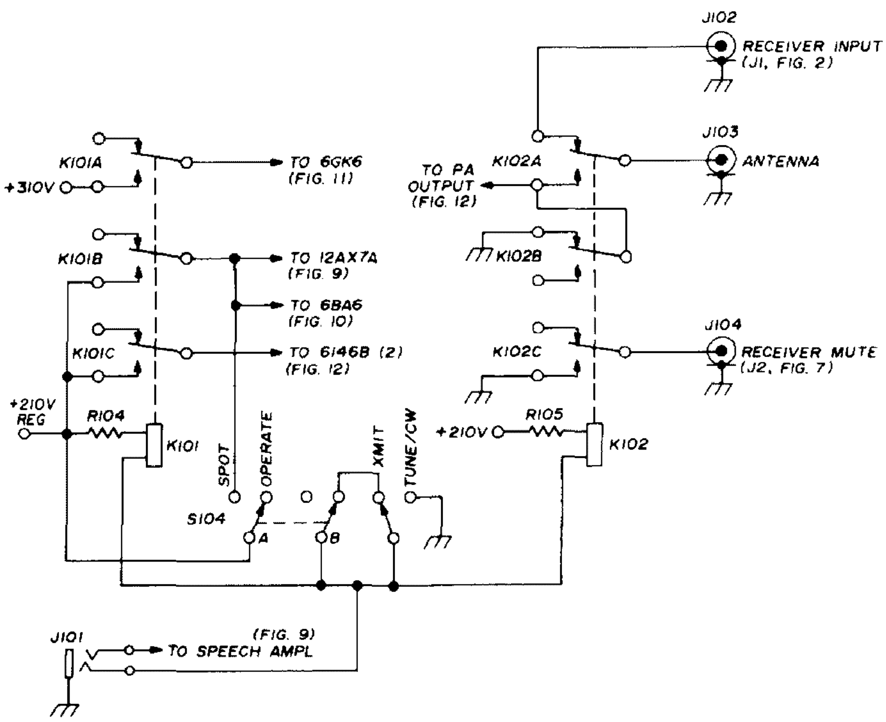


fig. 13. Voltage switching and transmitter and receiver control circuitry. Relays are shown in the standby (receive) position. Resistors R104 and R105 are voltage-dropping resistors, value chosen to reduce 210-volts dc to proper relay operating voltage.

measured output to a maximum of about 60 volts rf. Tuning C106 should further peak the driver output rf voltage.

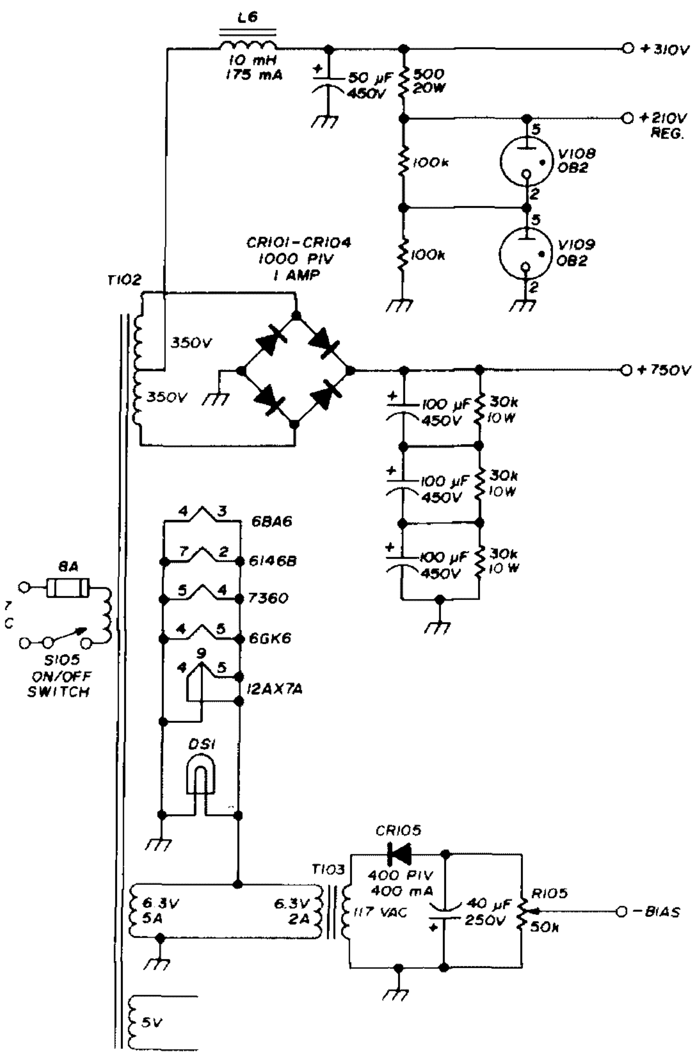
As a final check of proper alignment, the exciter's single-sideband signal should be monitored with an amateur-band receiver. Keep the 6146Bs out of their sockets during this testing. Place S101 in the *lower sideband* position, S104 in the *transmit* positions and switch S102 to *internal vfo.*

Keep the rf probe on the grids of V106 and V107 and loosely couple the receiver to L104. With the PTT switch closed and without modulation, virtually no rf voltage should register, indicating good carrier and vfo oscillator suppression. Speaking into the microphone should give readings of about 35 to 40

When adjusting the PA stage remember that more than 750 volts exists there, and if touched inadvertently, that's more than enough to prematurely end a promising career. Place the 6146B power tubes in their sockets and connect a high-wattage 50-ohm dummy load to J104. Adjust R105 to supply about -55 volts bias to the grids of V106 and V107. Set the drive control, R107, to about midrange and the relative power output sensitivity control, R106, to about one-third. All the other operating controls should be set as previously described when monitoring the exciter's signal, except that switch S101 should be in the *carrier/tune* position. With the PTT switch or S103 closed, adjust C107 and C108, the plate-tuning and antenna-loading capacitors, for maxi-

mum deflection of the relative power output meter, M101. Resistor R106, the output meter sensitivity control, should be adjusted to keep the meter at about two-thirds deflection at maximum power output.

Finally, insert a dc milliameter in the 750-volt line, and with S101 in the *lower-sideband* position and the PTT switch closed, adjust R107, if necessary, to bring the final amplifier idling current to about 60 mA. As an alternative, a



I101	6-volt lamp (power on)
L106	10-mH, 175-mA filter choke (Hammond 193J or similar)
R105	50k, 10-watt potentiometer (bias control)
T102	117-Vac power transformer, 350-0-350, 6.3 Vac secondary (Hammond 273BX or similar)
T103	117-Vac filament transformer, 6.3-Vac secondary (Hammond 166L6 or similar)

fig. 14. Power supply for the 40-meter ssb and cw transmitter.

50-ohm dummy load wattmeter or an in-line monitoring oscilloscope can be used to tune the PA stage and adjust the rf output meter. Capacitors C107 and C108 are then simply adjusted for maximum rf output as read on the wattmeter, or as seen by maximum deflection of the carrier envelope on the scope. R106 is then adjusted for two-thirds meter deflection at maximum power output.

operating

Operating in either the ssb or CW modes is extremely simple. For both ssb and CW transmission the transmitter is initially tuned up as previously desceibed for full carrier output. To use the exciter independently, S102 is switched to the *internal vfo.* With S101 in the *lower-sideband* position, placing S104 in the *spot* position will allow you to "talk" yourself on frequency while monitoring your signal with the station receiver. Once on frequency, simply switch S104 to *transmit* and all operating functions are controlled by the PTT switch.

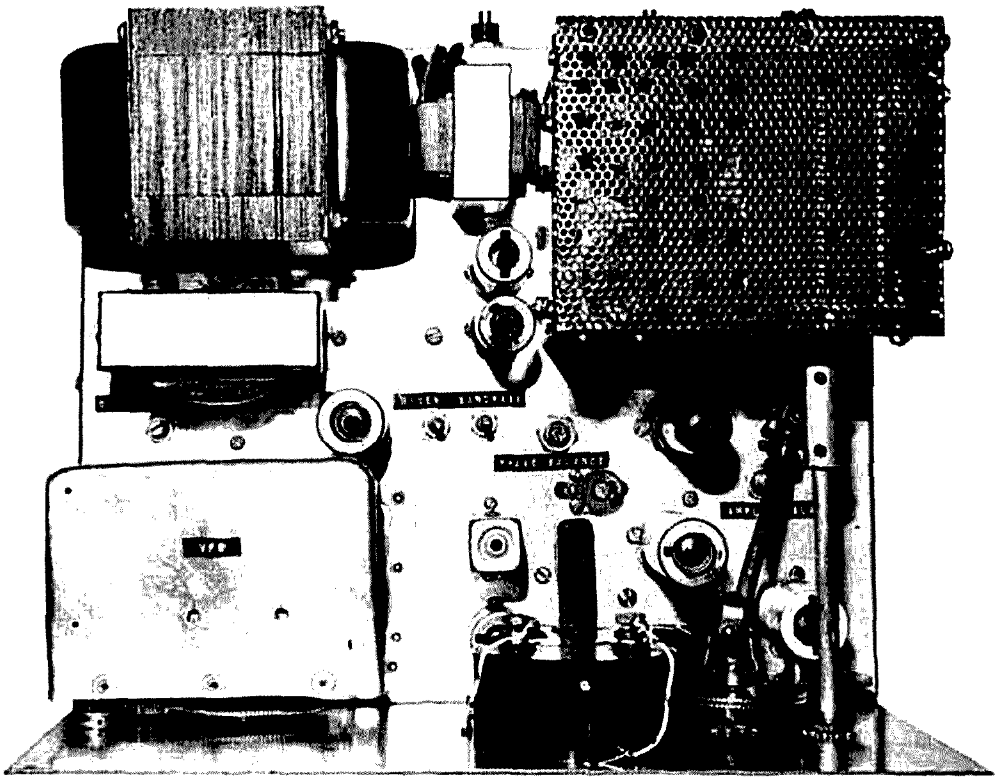
To operate in the transceive mode, connect the receiver vfo output to J105 using a shielded cable and switch S101 to the external vfo position. The exciter and receiver are now slaved to the receiver's vfo frequency, and ssb operation is the same as in the split function.

Although this exciter is primarily designed as an ssb unit, it puts out an excellent CW signal. To operate CW, simply tune up for maximum carrier output, then insert a key into J102. Switch S102 is left in the *carrier* position and S103 is used as a manual *CW/standby* switch.

circuit improvements

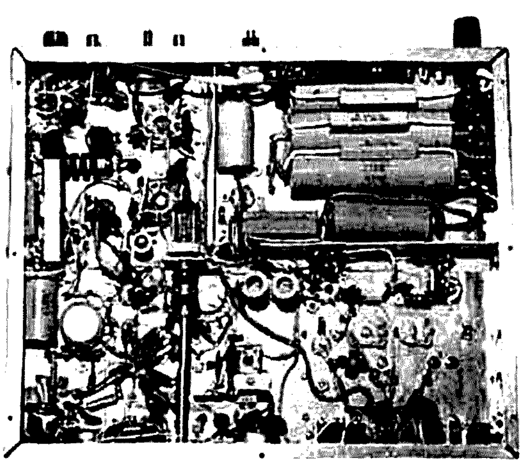
You may be surprised by the omission of some operating luxuries in this exciter. This was intentional – to keep design and alignment as simple as possible and to keep cost at a minimum. However, vox and sidetone circuitry could be easily added for *voice-operated ssb* operation and semi-break-in CW operation.

For those of you who may want to use upper sideband on 40 meters, switch



Top view of the transmitter chassis. Power amplifier cage is upper right, next to the power supply. Vfo enclosure is lower left, next to the front panel. Speech amplifier tube is underneath the antenna-loading capacitor shaft to the extreme right.

S101 could be changed to a 3-pole unit to switch in the required 453.75-kHz crystal. If you thrive on the sight of swinging meter needles, appropriate switching and metering could easily be added to monitor final plate current, plate voltage and grid current. And, for the few courageous builders who seek still more, multibanding this exciter might even be considered. Changing vfo frequencies and switching in separate mixer and driver coils would be the most simple means of adding 80- and 20-meter operation. For the higher frequencies, a second frequency conversion scheme would probably offer the best results.



Underneath the transmitter chassis. Front of chassis is at bottom. Power supply components are in upper right-hand corner.

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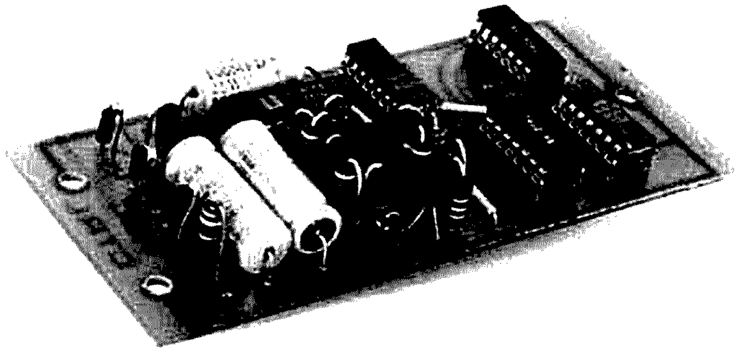


automatically controlled access to open repeaters

A control system that automatically guards against interference while allowing open repeater operation

R.B. Shreve, W8GRG, 2842 Winthrop Road, Cleveland, Ohio

Anyone trying to operate an open repeater on one of the popular frequency pairs such as 146.16/76 in an area like northern Ohio is increasingly faced with a difficult choice: guard the input, and shut out the visitor or occasional user who lacks the required means of access, or leave it unguarded and expose the repeater to a lot of plucking and other unwanted transmissions. Even if the input frequency is less popular, the growing availability of synthesizers puts an unguarded repeater at the mercy of the operator who can't resist seeing how many squelch tails he can generate and how many identifiers he can trigger each time he pushes the button on his microphone.

WR8ABC, the 146.16/76 repeater serving the Cleveland area, is located on the crest of a ridge southeast of the city. There are 16/76 repeaters to the east, at Ashtabula, to the south at Newcomerstown and Columbus, and across Lake Erie in Detroit. Even without a band opening, mobiles in the fringe areas between repeaters often key more than the machine – with only a slight propagation enhancement the situation can become chaotic.

The Detroit repeaters, Great Lakes on 146.16/76, and DART on 146.04/64, have adopted sub-audible tone access, or full-time private-line. The Ohio 16/76 repeaters have established secondary inputs on discrete tertiary frequencies (for

example, 146.355 in Cleveland) to permit base stations and high-powered mobiles in the fringe areas to select the repeater they want to access.

the problem

The group that operates WR8ABC has consistently voted to keep the repeater as open as possible. When conditions made it necessary to guard the 146.16-MHz input, provision was made for access with made it possible for a fringe area station to picket-fence the repeater indefinitely as long as his signal was strong enough to open the receiver squelch once every five seconds, which was hard on the equipment and even harder on the control operator. More than once the repeater has been out of service for several hours because a control operator switched it off and forgot to turn it on again when the interfering signal disappeared.

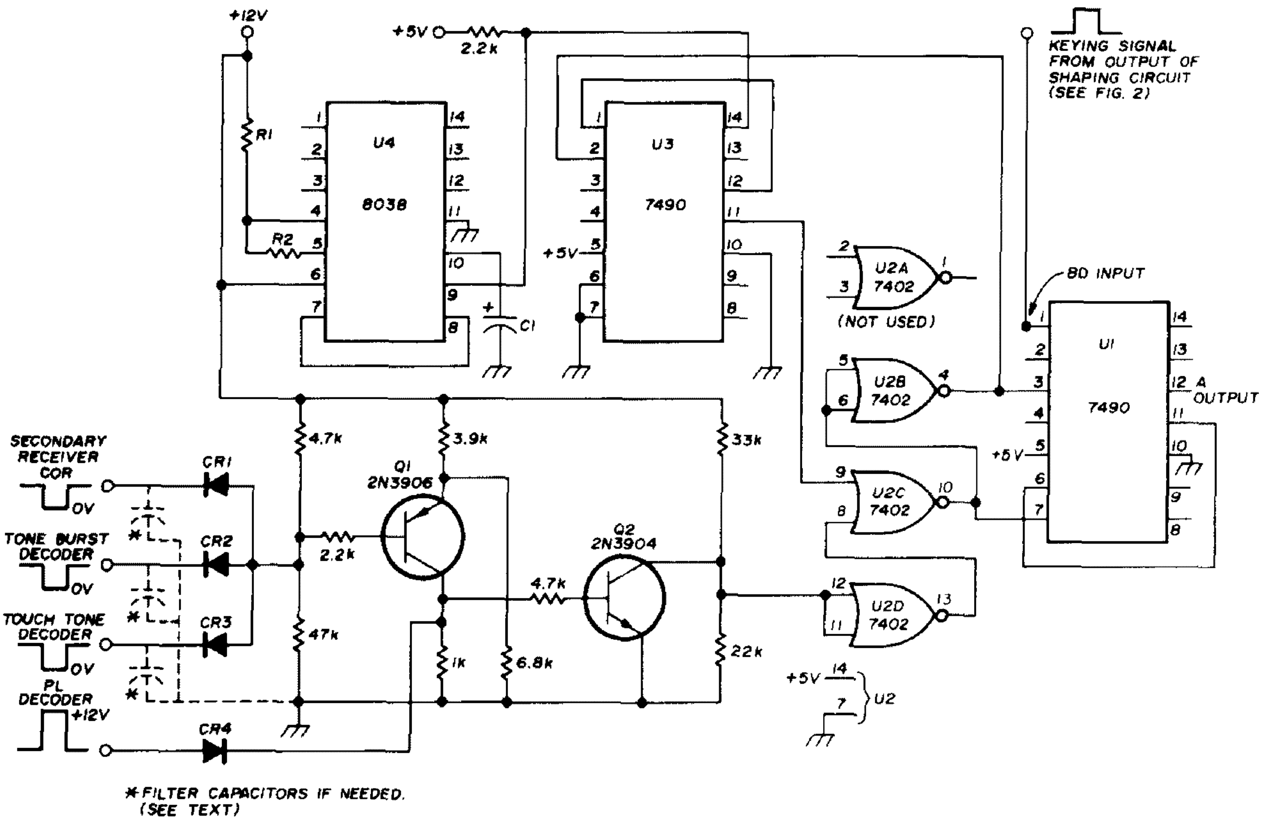


fig. 1. Timing circuit and counter for the repeater control system. For values of R1, R2 and C1, see text. All diodes are general-purpose silicon types such as the 1N914.

private-line (110.9 Hz), tone burst (2000 Hz) and the 1336-Hz tone generated by a Touch-Tone pad. In addition, once the guard was opened by one of these means, the repeater remained open to access by any on-frequency carrier for the duration of each transmission and five seconds thereafter.

Permitting "tail-ending" in this way effectively opened the machine to use by any number of stations, with or without private-line or tone generators, as long as one of them could whistle the machine up in the first place. Unfortunately it also

the solution

In an attempt to overcome some of the disadvantages of the previous method, a new system to control access to the 146.16-MHz input has been installed which leaves the input open until it is repeatedly keyed by a signal which is not modulated by any of the access tones recognized by the decoder. When the receiver squelch is operated three times in succession by such a signal the input is automatically guarded. The guard is opened, also automatically, by a timer in approximately 15 minutes; it opens

sooner if the repeater is keyed by a station using an accepted access tone, or by one using the unguarded secondary input.

The circuit is shown in fig. 1. The BD input of an SN7490 decade counter, U1, is toggled by a signal source that goes high each time the guarded receiver COR

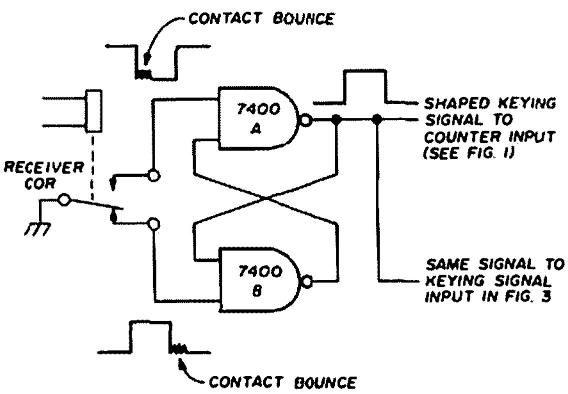


fig. 2. Shaping circuit for the keying signal.

is triggered by a signal. It counts the pulses generated by a signal with no access tone, and latches in the "binary 9" condition on the fourth count. This generates a logic 1 at the A output, pin 12, until the counter is reset to zero.

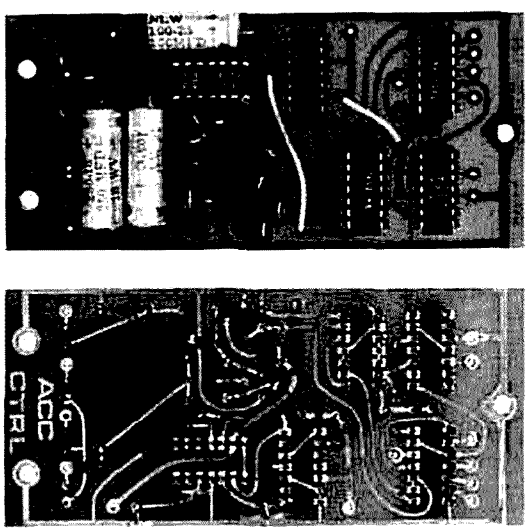
To make the counter operate in this manner the D output, pin 11, is wired to one of the R_9 reset inputs, and the other R_9 input and one $R_{\emptyset}$ input are controlled by an SN7402 NOR gate, U2. So long as one $R_{\emptyset}$ and one R_9 input are low, U1 counts the receiver COR cycles. On the fourth count the D output at pin 11 carries the R_9 input at pin 6 high. If pin 7 is also high, the counter latches with both A and D outputs in the high state. It will remain in this condition until the second R_9 input, pin 7, is switched to the low state by gate U2C. The same signal that does this switches the output of gate U2B high and U1 resets to zero.

The impulse to reset the counter to zero can come from any one of the four sources connected to the resistor-transistor network through diodes CR1 through CR4. The private-line (PL) output from the WR8ABC PL decoder is low when the receiver is squelched and high when the squelch is opened by a private-line tone. The inputs to the other three diodes are normally high, and are switched to ground potential when an activating signal is received.

If no signal is received from any of the four sources connected to the diodes for a predetermined interval, the timer connected to the second input of gate U2C will reset the output counter. The timer is another decade counter, U3, driven by the square-wave output of an Intersil 8038 precision waveform generator, U4. The rate at which the 8038 cycles is controlled by the values of C1, R1 and R2. As installed at WR8ABC, C1 = 100 μF, R1 = 47k, and R2 = 100 ohms. The generator cycle is just under two minutes, and U3 reaches the count of 8 in 15 minutes, at which time it resets itself and U1.

construction

Note that the 2.2k load resistor connected to pin 9 of the 8038 is returned to +5 volts rather than to +12 volts to make the square-wave output TTL compatible. Some precautions in construction are advisable. Rf shielding is essential if the unit is to be installed near a transmitter, although there should be no problem with a split-site repeater. The count and reset inputs are all sensitive to short pulses. Signals connected to the PL and



Front and rear views of the printed-circuit board used for the repeater access system. The relay is not mounted on the board.

tone inputs that reset the unit should be filtered to suppress response to transients, and so should the signal from the secondary receiver COR. A small electrolytic capacitor or a resistor and capacitor should suffice.

The keying signal to the counter input is more critical. COR relay contact bounce can cause erratic counting. The best way I have found to shape the COR signal is shown in fig. 2. This system uses two sections of an SN7400 NAND gate and a double-throw COR relay on the receiver.

The unit was designed to work with the solid-state control logic described in an earlier article,[1] but the output can be used to drive any TTL-compatible logic, or to operate a transmitter keying relay with a transistor driver. A circuit to do this using the other two sections of the SN7400 is shown in fig. 3. Transistor Q3 is a pnp silicon switching transistor capable of handling the keying relay current and voltage. Resistor R3 should be selected to limit the current through it to approximately 10 mA when the junction of R3 and R4 is at ground potential.

Resistor R4 should limit the voltage at this junction to between 3.5 and 5 volts when the junction is ungrounded. Diode CR5 protects Q3 against surges when the relay opens. The *access control on/off* switch, which can be a remotely controlled flip-flop or relay, permits the operator to inactivate the guard circuit if a completely open operation is desired at any time.

The original unit in service at the WR8ABC repeater is built on a homemade PC board approximately 2x4 inches, which provides for all the components in **figs. 1** and **2** except the COR relay. Boards for the later model shown in the photographs, which includes the optional input filters, and/or complete assembled units are available from the author.*

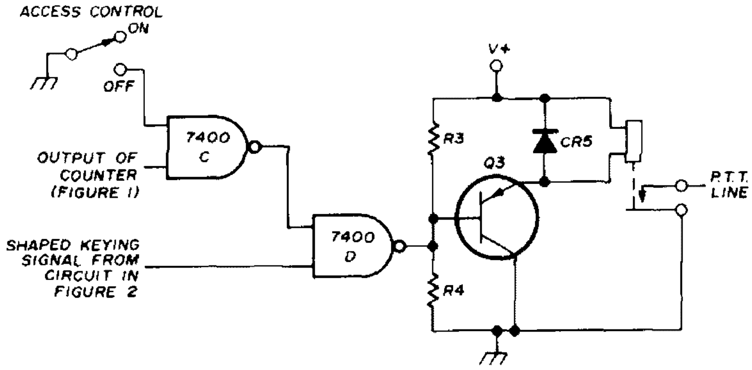


fig. 3. Transmitter control relay circuit. Values of R3 and R4 are selected as described in text. Transistor Q3 is a pnp silicon transistor capable of switching the relay current and voltage. Diode CR5 must have a PIV rating higher than the positive voltage supply.

conclusion

To summarize operation of a repeater with this system of access control, so long as no more than three successive signals without access tones key the transmitter through the guarded receiver in any 15-minute period, the guard will remain open indefinitely. It will close immediately, however, when the guarded input is plucked three times, accidentally or on purpose. A visitor to the area, without tone access, never has to wait more than 15 minutes for the repeater to open up, and since he is usually answered by a station using PL, tone, or the secondary receiver, can carry on a conversation indefinitely in most instances as though the repeater was unguarded.

reference

1. R.B. Shreve, W8GRG, "Integrated-Circuit Sequential Switching for Touchtone Repeater Control," *ham radio,* June, 1972, page 22.

ham radio

*Epoxy printed-circuit boards can be purchased from the author for $5.00, postpaid. Prices on complete units, with any reset time interval up to 30 minutes, and designed to interconnect with the user's equipment, will be quoted on request.

six-meter frequency synthesizer

Complete construction details for a frequency synthesizer that covers the entire 50-MHz band in 1-kHz steps

Kenneth W. Robbins, W1KNI, Sperry Research Center, Sudbury, Massachusetts

Although I have been relatively inactive on 6 meters in recent years, this band has always been a favorite. This is no doubt due to an austere but exciting ham beginning on 5 meters during the 1930s when modulated oscillators, superregen receivers and Pickard antennas were common. After World War II modulated oscillators were phased out by stability regulations, and crystal control became mandatory.

Rock-bound transmitters forced development of improved vacuum-tube vfos over the years, and these gradually gave way to solid-state circuits. Heterodyne versions improved on them in turn and now, with the "galloping IC Technology," synthesizer frequency control is coming on strong. Widely used in military and commercial equipment for many years, until recently frequency synthesis has been too costly for general ham use. However, two-meter units are now on the market and a number of construction articles have appeared in print for the do-it-yourselfers.

Recently, I had a go at working up one of these exotic channelizing vfos for a 50-MHz a-m rig. New and unusual circuit problems had to be solved before success was attained. For the unwary homebuilder who is or will be building an indirect synthesizer, this article will endeavor to point out some constructional pitfalls, offer a few guidelines and, hopefully, aid in maximizing proper operation on your first try.

design

Following a modest literature search, a 6-meter IC frequency synthesizer was blocked out that tuned from 50.000 to 54.000 MHz in 1.0-kHz steps. Considera-

tion was given to minimum steps of 5 or 10 kHz, knowing it would ease the task of phase-locking and reduction of spurious outputs, but this less desirable trade-off was finally overcome by intensive debugging. To be useful as an 8-MHz crystal replacement, output frequency runs from 8,333,333 to 9,000,000 Hz. An electronic calculator made quick work

enameled wire, a handy enameled hookup wire with insulation which melts back from hot solder to leave a nicely tinned end ready for connection. Feeding in clock pulses having a configuration called for by the chip manufacturer at a 8,333,330 Hz rate resulted in an output that closely resembled a pseudo-random bit stream! "Oh well!" said I, "probably

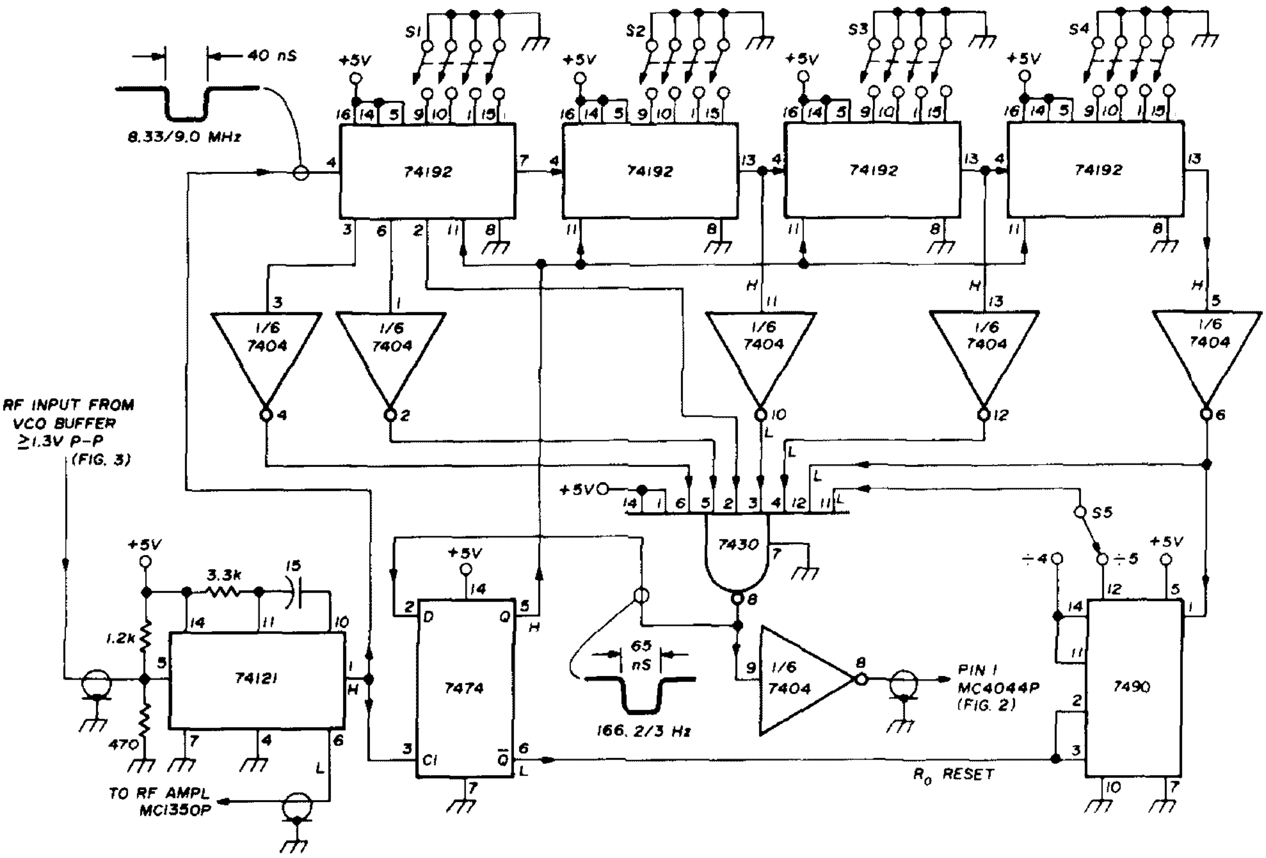


fig. 1. Programmable divider divides from 50,000 to 54,000 in steps of 1. The thumbwheel switch assembly (S1, S2, S3 and S4) is an EccoSwitch type 4R177612G. Switch S5 provides divide-by-4 for MARS netting (see text).

of the math. Frequency increments of 166.2/3 Hz, when multiplied by 6, yield 1.0-kHz steps on 6 meters; the required division is 50,000 to 54,000. A 1.0-MHz reference crystal divided by 6000 produced the necessary 166.2/3 Hz comparison signal for phase detection.

First off, a string of five 74192 programmable divider ICs were plugged into sockets Duco cemented to perfboard and wired to divide by 50,000 as shown in fig. 1. This wiring consisted of no. 28 Beldsol

just a wiring error, maybe a stray oscillation or a bit more B+ bypassing."

Three weeks later, the output wasn't quite so random but counting down to zero from a given input frequency never agreed with the divider truth table. Even getting close to 166 Hz required a higher input frequency than theory said was necessary. The B+ supply from a three-terminal IC regulator *had* to exhibit very low impedance. Miniature axial-lead tantalum capacitors added for supply voltage

bypassing helped a little, but it was just impossible to obtain a stable, fixed division ratio.

About this time K2OAW described his programmable divider.[1] After some reading, the operating principles of K2OAW's circuit became clear. Since I only had transmitter frequency control in mind, his circuitry for receiver LO use could be deleted. Also, since the most significant digit never changed, a 7490 wired biquinary could divide by five.

This new circuit replaced the first version and at least got me into the ballpark, meaning division was off by only scores of numbers instead of hundreds! Many troubleshooting tricks were tried and failed. Finally, a fruitful discussion with WA1CTS yielded a nugget generally known to computer builders but seemingly little known among amateurs: B+ and ground for TTL logic should use ground-plane techniques with a 100-ohm transmission line built in for power distribution. Neglect of this basic design requirement was manifested in extra counts. My skinny little magnet wire must have looked like kilohms!

Not wishing to process double-clad printed-circuit board, and with low cost always in mind, the socket/perfboard layout was modified to accomodate 3M's 1181 half-inch wide, copper-foil adhesive tape, one pattern stuck down for ground and the same pattern directly opposite for B+. This formed a low dc resistance, low inductance and distributed bypass capacitance power feed system that also made socket wiring easier. When fired up once again, the long sought after magic numbers appeared!

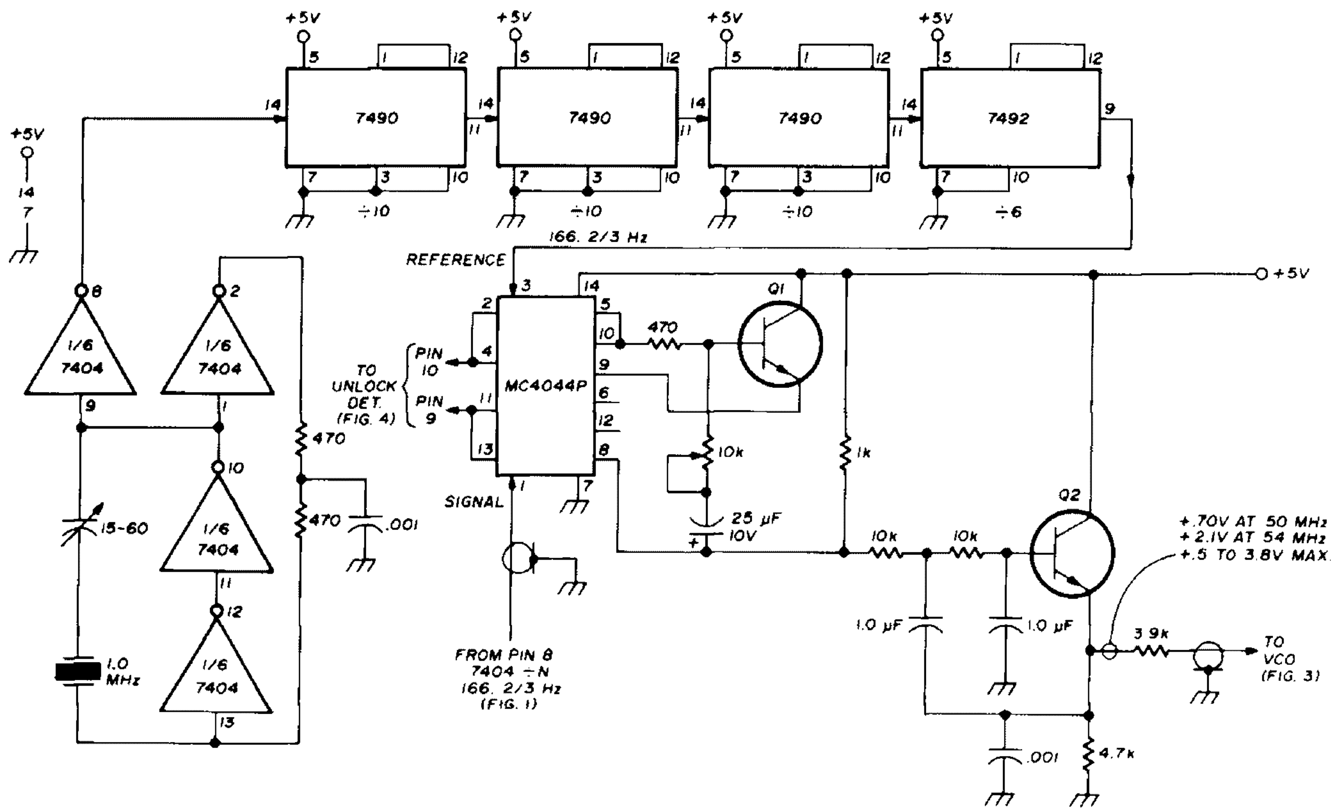


fig. 2. Circuit for the 1-MHz reference oscillator, divide-by-6000 and MC4044P phase detector. Transistors Q1 and Q2 may be any small npn transistors with current gain of at least 100 such as the 2N4124 or 2N5130.

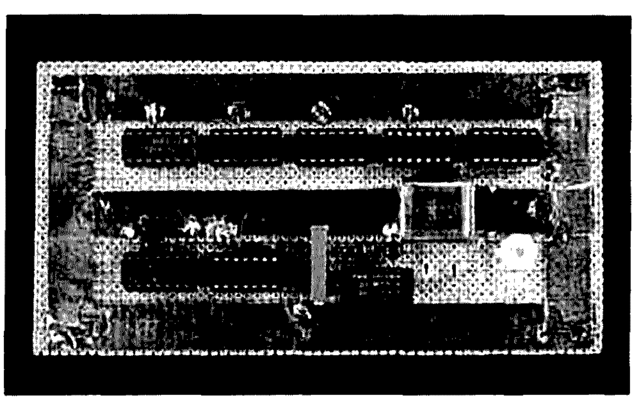
Dividing 8,333,330 by 50,000 gave 166.6 Hz, and dividing 9,000,000 by 54,000 also gave 166.6 Hz. Various in-between divisions were tried; all came out correctly. This without *any* B+ bypassing to ground, too. Success was so sweet, the foil transmission line impedance never did get measured. However, 4.2 pF per half square inch of foil was measured and with 20 inches per board this is 168 pF. As a rule of thumb,

pulse rise time can be related to one cycle of rf in the same period, so the 5-nanosecond transitions correspond to 200 MHz where 168 pF has a reactance just under 5 ohms. Obviously, this is quite effective in shunting pulses whose source impedance is 10 to 20 times higher.

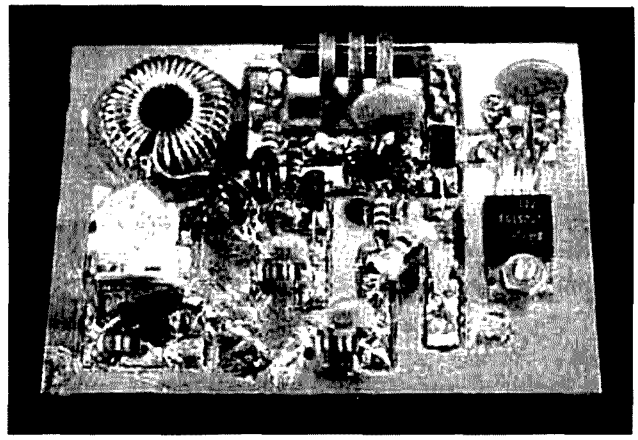
With an idealized clock pulse obtained from a lab pulse generator the next item of business was to simulate it inexpensively. A 74121 one-shot, timed to deliver 40-ns negative-going pulses, was found to be perfectly adequate. Proceeding next to the reference circuit and again using copper foil for power distribution, a 1.0-MHz crystal oscillator divided by 6000 was constructed (fig. 2) and never failed to operate properly from first turn-on.

vco

As other writers have indicated, the vco must be inherently stable by itself before electronic frequency control is fed back. A Vackar oscillator circuit was chosen for three reasons: it is frequency and amplitude stable, and, like the Clapp, has one connection where small capacitance variations result in large frequency shifts. The circuit, fig. 3, was built on a bit of double-clad printed-circuit stock to insure mechanical integrity and simplify shielding. It was then tested by feeding in varactor bias from a 10-turn 10k pot tied across 5 volts of B+. Tuned-circuit component values were juggled to obtain the



Layout of the reference divider, showing the use of 3M adhesive-backed copper tape to achieve the low-impedance B+ and ground leads required for proper operation of the TTL logic. Circuit board for the programmable divider (fig. 1) uses the same technique.



Small size of the circuit boards used in this synthesizer results in a very compact package.

graphed tuning curve shown in fig. 4. This check should always be made to get an idea of linearity and voltage swing required for a given configuration.

When tuned in on a CW receiver, the carrier jumped in discrete steps as the pot wiper moved along individual turns of winding resistance which is understandable when it is realized that slew sensitivity is about 500 Hz per millivolt. Slow frequency drift is not important during this test but any audio rate burble must be eliminated. Some rectifier diodes work well as tuning devices but be careful of ambient light effects; one diode I tried had a translucent plastic case which caused 60 Hz fm from an overhead light until the device was wrapped in black plastic tape.

Oscillator B+ must be essentially battery-pure and stable, and a separate 7805 IC voltage regulator is recommended. No doubt a μA723 regulator could be used, but these new three-terminal voltage regulators are so easy to wire in, they spoil you for anything else.

Unilateral amplification is necessary to prevent spurious vco pulling by logic pulses sneaking back through the gain chain. The resultant pulsed fm generates a wide spectrum of hash that is impossible to cure without redesign. A Darlington emitter follower has worked well in this regard and is able to properly fire the one-shot. Additional isolation and gain is provided by a MC1350P which drives the transmitter multiplier.

phase detector

Next on the agenda was selecting and optimizing a phase detector circuit. Perhaps it would be helpful to set up an idealized specification and then see what can be done to meet it in the real world. This black box would detect the slightest difference in phase of incoming 166-Hz pulses, change this information into a precise step of dc control voltage and instantly slew the vfo back into exact synchronism with its reference. In truth, since nothing works in zero time, there would have to be *some* phase difference to produce a correction voltage, and steps of dc voltage generate wideband transients. Therefore, a real world circuit will necessarily have time delays, small error signals and low pass filtering as minimum requirements.

The first circuit I tried[2] was totally unsuited for low-frequency use; the vco put out a spectrum of hash many hundreds of kHz wide and optimization only reduced this to about 90 kHz. For audio-frequency phase detectors, many designers have gone to sample-and-hold circuits as a means of reducing pulse feedthrough, but since I didn't have any enhancement-mode fets on hand, a pair of MPSA12 Darlington transistors were substituted during a second effort. After considerable playing around with component values, sidebands were down to about 20 kHz. One bad feature was the lack of drive toward sync if the vco happened to get outside one end of lock range. For all its faults, a MC4044 IC always drove towards lock, no matter where the vco was initially.

Since W1UYK's circuit[3] worked on 41-Hz pulses I decided to give it a try. It was wired in per his schematic and showed promise right away; sidebands came down to about 10 kHz. Another period of testing commenced in an attempt to modify the circuit for this particular synthesizer. The final result is similar with one interesting exception — the use of a 10k variable for adjusting loop lock-up rate.

As resistance is progressively reduced, vco slew response changes from over to critical to under damping and eventually into sustained hunting. It's quite easy to hear this effect on a CW receiver and adjust for a rapid settling time (about 2 seconds) by placing a finger momentarily on the vco tank coil to force it far out of lock. Do this at 8.33 MHz where loop gain is highest.

Settling time is a little longer at 9 MHz

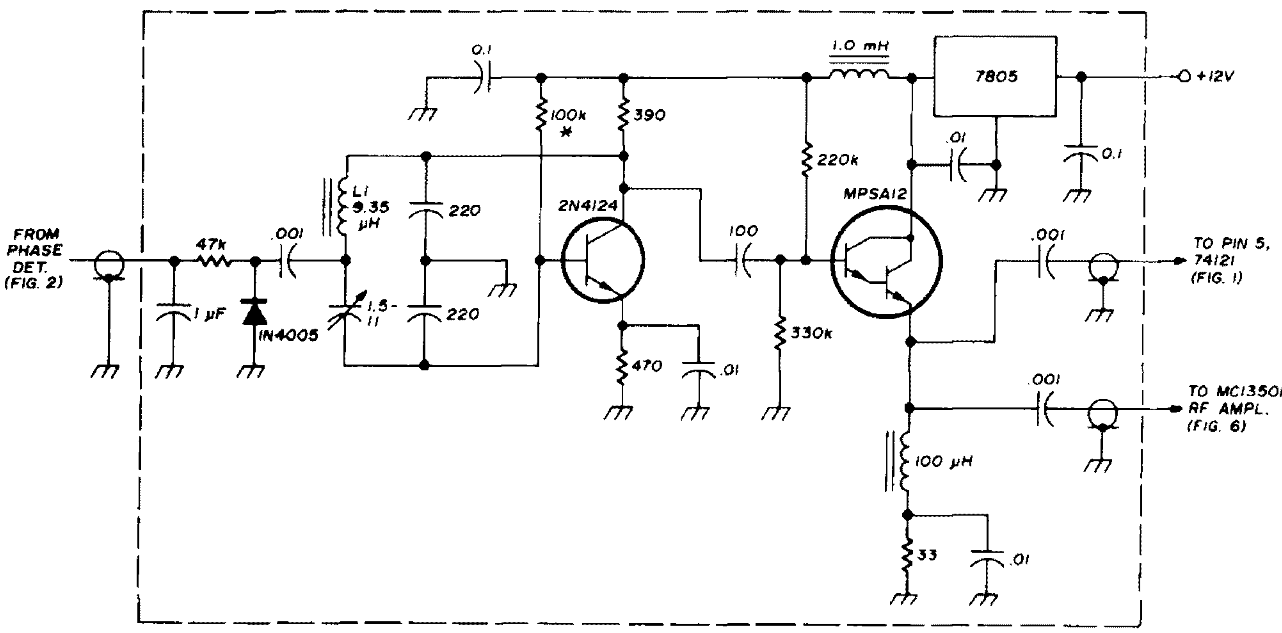


fig. 3. Circuit for the vco used in the 6-meter frequency synthesizer. The 9.35 µH inductor, L1, consists of 37 turns no. 22 enamelled on a T-68-2 Amidon toroid core.

due to slightly lower voltage sensitivity and is a tradeoff made for *non-linear* varicap pull range. A low leakage electrolytic must be used for the 25-μF capacitor (20 megohms on a Simpson 260) because it connects between a possible maximum 4 volts and a dc amplifier having a current gain of about 50,000. *Only* ac coupling is wanted. An RC filter follows the active filter to reduce sidebands to essentially zero and at 40 dB over S9, the carrier sounds perfectly quiet.

In a lab check using a special oscilloscope, varicap control voltage showed 10 microvolts pulse and about 30 microvolts random dc on a plus 1.0-volt pedestal at 8.5 MHz.

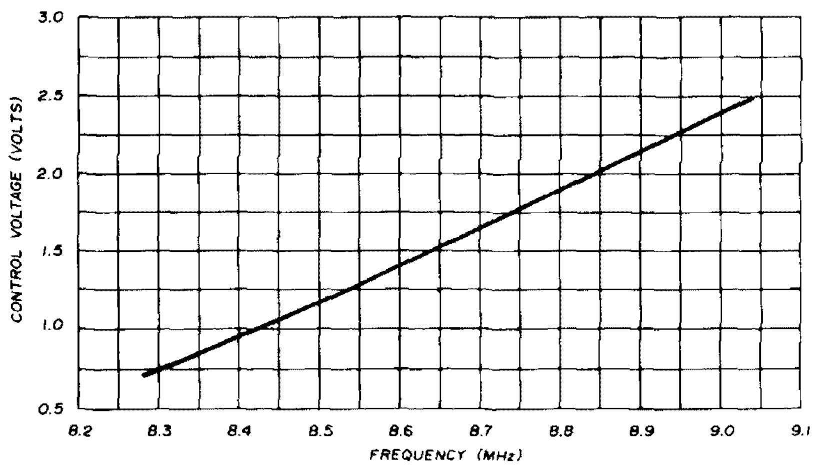


fig. 4. Frequency of vco vs control voltage, using a Sylvania 1N4005 rectifier as a varactor.

This equates to an average carrier uncertainty of ±7 Hz or six times that on 6 meters, an acceptable figure for most transmission modes. These small error voltages remain unchanged, but the major dc voltage will lie between 0.5 and 3.8 volts, depending on frequency and trimmer capacitor adjustment (0.7 at 50 MHz and 2.1 at 54 MHz used here).

A simple but revealing test can be made by tuning in the reference crystal's 9th harmonic at 9.0 MHz, then setting in a division of 53,999 to produce a 166-Hz heterodyne. Servo loop *limitations will be* evidenced by a small burble or beat note instability. Any circuit modifications should be aimed at minimizing this randomness without degrading sideband levels or lock-up action.

unlock detector

If the synthesizer becomes unlocked for any reason, transmission of off-frequency signals could take place. To prevent this, a simple pulse-width comparator, timer and relay deactivate the transmitter and hold it off until sync is regained (see fig. 5). Two NAND gates make up a one-shot whose timing capacitor produces one microsecond for every 770 pF, so .01 μF gives about 12 microseconds. This is longer than the normal up-to-5 microseconds out of the first gate. These two signals take turns keeping inputs low on a third gate and its output stays high. Should the vco suddenly shift frequency more than 10 kHz at 6 meters, the 5-microsecond pulse briefly becomes longer than the 12, gate output drops, firing the timer and energizing the relay for 4 seconds. Consequently, a frequency change of several kHz is possible without initiating action.

A desired frequency shift tolerance may be set in by variation of the one-shot capacitor while off time is changed by appropriate shift of R or C in the NE555 timer. Start-up, long settling times or hunting will keep the relay on; it will not reset until phase lock is effected.

construction

If you decide to build this six-meter frequency synthesizer, there are several important points to keep in mind. First of all, use a ground plane or copper strip for logic B+ and ground power distribution. Strive for battery-pure dc supplies for the *oscillator and buffer*. Use at least

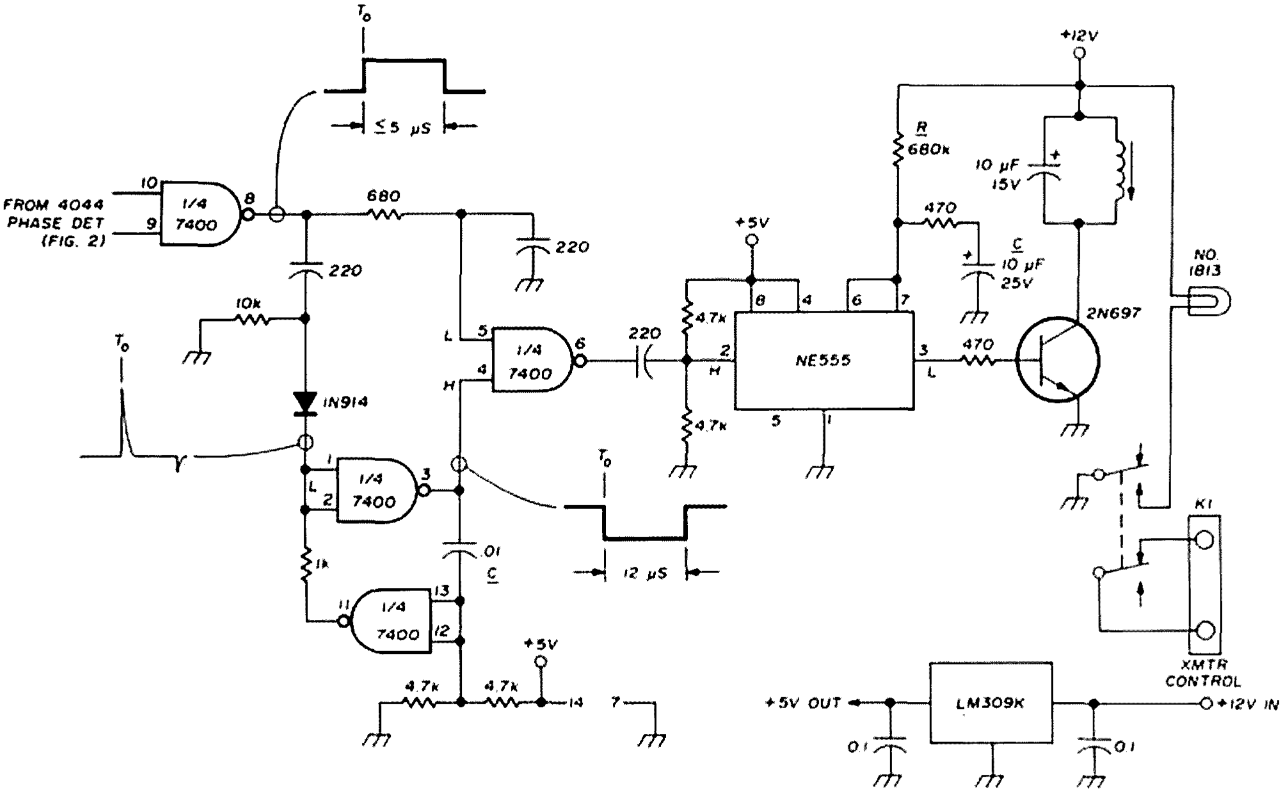


fig. 5. Circuit of the unlock detector and timer. Relay K1 is a Sigma 62R2-12DC. Simple three-terminal five-volt voltage regulator for the entire synthesizer is shown at right, below.

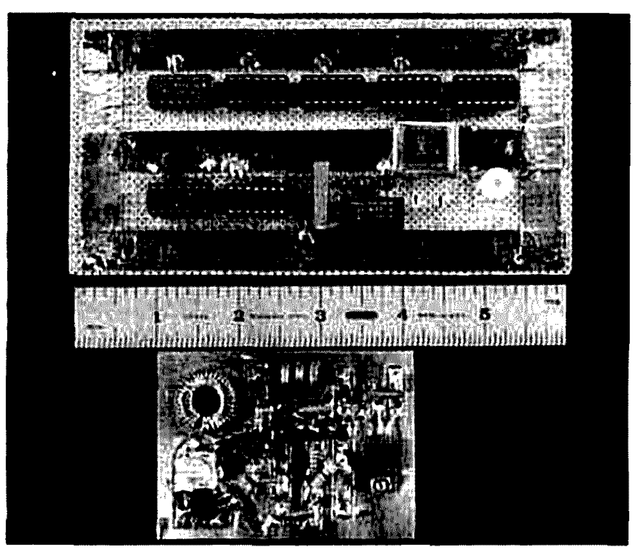
two voltage regulators: One for logic, one for the vco. Zener diodes have too high impedance and unregulated B+ is out of the question.

Keep the dialed logic wiring short and direct to its respective IC terminal. Optimize the base-bias resistor in the vco using the 9-MHz burble test (use a 22k isolating resistor and 250k pot in series to determine the proper value). Follow the vco with one-way rf amplification.

Shield the entire synthesizer. Shield the oscillator separately, and use feed-through capacitors and shielded cable. Beware of ground loops – non-reducible sidebands usually result. All construction must be mechanically secure. Anything that moves will cause phase shifts that the detector tries to correct for. Think of phase as a change of less than one-half Hz at 8 MHz!

Stray capacitance at the varicap connection will have considerable affect on the capacitor values required to bracket 8.33/9.00 MHz with a given tuning diode and bias voltage swing. A frequency counter is convenient, but a well-calibrated communications receiver will do fine. If only the first MHz on six meters is of interest, adjustment is that much easier. Experiment with different voltage-variable capacitors. A Sylvania 1N4005 rectifier diode produced less frequency jitter than a Motorola Epicap MV2209.

For a real eye-opener, try placing a battery-powered broadcast radio next to the programmable divider board to pick up its amazing spectrum of signals. Then,



Construction of the vco. The 9.35-μH inductor is in the upper left-hand corner of the board.

"finger test" the vco – the result is hard to believe!

summary

Troubleshooting this synthesizer was a real challenge but definitely worthwhile once it began to operate correctly. There is tremendous satisfaction in having a 6-meter crystal-stable rf generator with 4000 discrete frequencies. Great for nets, receiver calibration and avoiding interference. When asked to move "up a couple," you'll shift *exactly* two kHz. Schedules on a prearranged frequency will be right on. Other vfos can be

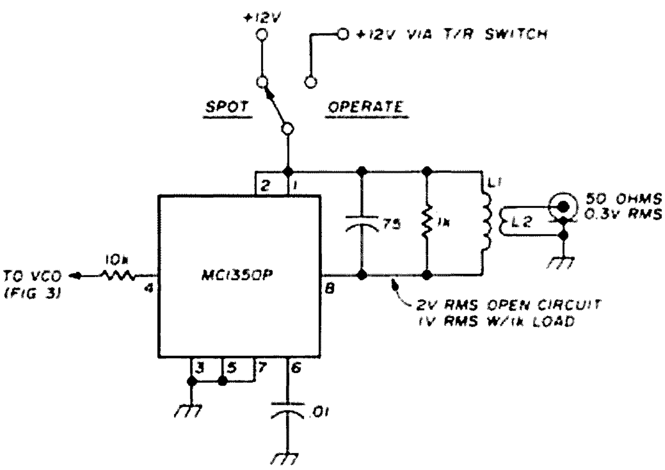


fig. 6. MC1350P rf amplifier provides 0.3-volt rms output into 50 ohms. L1 is 19 turns no. 28 on 3/6" diameter iron-core form, 3/4" long. L2 is 6 turns no. 28 over center of L1.

calibrated by your dialing in spot frequencies to zero in on. MARS netting is possible by adding a toggle switch on the 7490 bi-quinary to divide by four; see the schematic in fig. 4. Narrow-band fm is possible by adding modulation to the vco control voltage, but hum will be a problem as only tens of microvolts can be tolerated.

references

1. P.A. Stark, K2OAW, "Frequency Synthesizer for Two-Meter FM," *73*, October, 1972, page 15.
2. K.W. Robbins, W1KNI, "Tunable Six- and Ten-Meter Phased-Locked Loop," *ham radio*, January, 1973, page 40.
3. D.H. Stevens, W1UYK, "A 4000-Channel Two-Meter Synthesizer," *QST*, September, 1972, page 17.

ham radio

performance characteristics of vertical antennas

A discussion of matching networks, network losses and bandwidth of vertical antennas of various heights

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

I've been interested in antennas for 80-meter DX that were simple, inexpensive and effective, so I've been reading a lot of the vertical antenna literature prior to putting up a vertical or a phased vertical array. I chose verticals early in this effort since antenna books show that the low angle radiation from vertical monopole antennas is considerably better than from horizontal dipoles unless the dipoles are unreasonably high, at least for 80 meters.

The first question was, "What vertical height should be used?" A recent article[1] shows that short verticals do a pretty good job, which I agree with. Although some antenna articles have indicated that tall verticals have much better low angle radiation than short ones, reference to antenna books such as those written by Kraus[2] or Jordon[3] show that the low-angle radiation pattern is essentially the same for a wide range of vertical heights. More on this later.

My studies show that earth and network losses are the most important factors. These losses are greater for short verticals than they are for tall verticals.

I've indicated a few considerations, but there are still a number of questions to consider and answer, and in more detail. How high should the vertical be? Is a tall vertical better than a short one? If so, how much better? What does better mean, or what factors or tradeoffs are important, and what is their relative importance?

You will find that there is quite a bit of choice in the height that may be used for a single vertical. If two or more verticals are to be part of a phased array, then there is less choice as to which antenna height may be used.

Some of the factors that must be considered are self-impedance; mutual impedance in arrays; earth, radial, and network losses; bandwidth and vswr versus height; vertical radiation pattern; type of tuning network required; type of transmission line used; physical or mechanical factors; and the radial ground system.

How can you make sense out of so many interrelated factors? You don't want to reinvent the wheel, so I will make use of material such as that from early

issues (1930s) of the *Proceedings of the IRE*, and from standard antenna textbooks. Today there are a number of new tools at our disposal such as digital computers, pocket electronic calculators, etc., and these were used to develop answers to some of the questions.

Rather than try to answer all of the questions at once, several articles are planned, with only a few topics in each. The data and examples will cover a range of values which are practical for average amateur situations. The graphs will be large enough to serve as useful reference data for amateurs who wish to work out other examples.

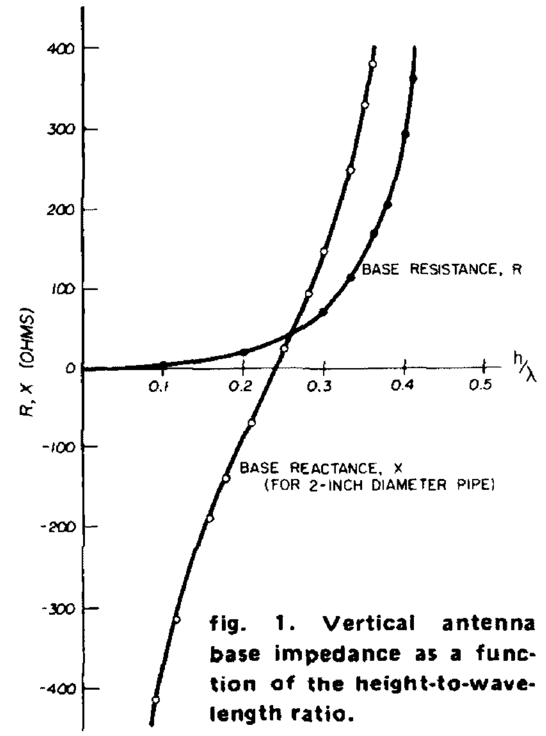


fig. 1. Vertical antenna base impedance as a function of the height-to-wavelength ratio.

This first article will deal with the self impedance of a single vertical, and indicate what networks (if any) are required for various heights of verticals, what losses occur in these networks, and what bandwidths result.

The second article will compare the vertical radiation patterns of various height verticals. The third article will discuss ground losses, and answer the remaining questions posed here. The last article will present mutual impedance data, and will analyze a particular phased array.

antenna self impedance

The electromagnetic field produced by an antenna results from certain current distributions on the antenna which vary with time. Many cases are analyzed in the literature, from the elementary current element, to the elementary dipole, to a full-length dipole with sinusoidal current distribution. Various steps in the analysis show that induction, radiation and electrostatic fields result, all of which fall off differently with distance. All of these fields (the near field) must be considered to determine either the reactance of a single antenna, or the mutual impedance between two or more antennas.

The radiation field is sufficient when considering the radiation patterns. These theories were used by authors of antenna books and articles to develop equations used to make calculations here. I used the appropriate formula from Jordan[2] for antenna self impedance (one for resistance and one for reactance). Each formula is a long complex expression consisting of many sine and cosine integral terms, and they are also functions of antenna height.

Antenna self impedance, as used here, consists of the antenna base resistance, R, and the antenna base reactance, X. You must be careful when using such formulae to be sure that loop or base values are used consistently and properly. The base impedance values are those seen at the

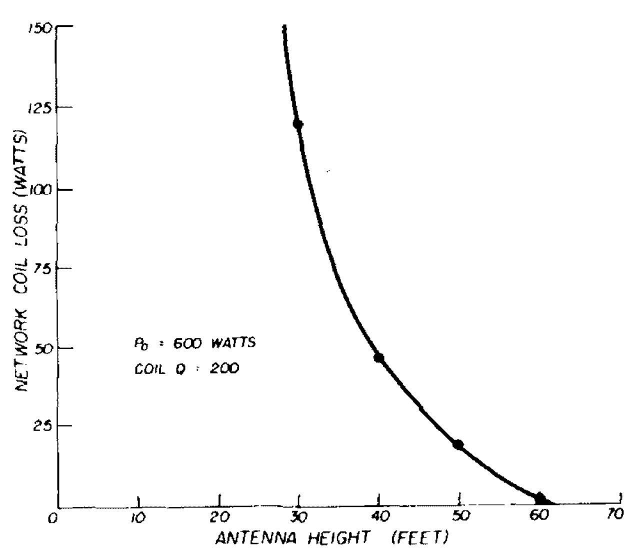


fig. 2. Network coil loss vs antenna height (see text).

base of a vertical antenna. Loop values result from considering the current loop up on the antenna. It is easy to convert from one set of values to another, as for example

$$R_{base} = R_{loop}/\sin^2(\beta h)$$

Antenna textbook data is sometimes given for loop values, and sometimes for base values. Base values are used here.

$$\beta = \frac{2\pi}{\lambda}$$

where λ is the wavelength, and h is the electrical length of the antenna in wavelengths.

antenna height

Another time you must be careful is when you calculate the physical length of the antenna. For antennas of the diameter of interest (2-inch diameter conduit was used here), the physical length should be 5% less than the electrical length. The computer calculations were made using the electrical length in order to compare my results with standard antenna textbook data, but the physical lengths given here will have the 5% shortening included. A handy formula for this is

$$HA = \frac{984 \times h/\lambda}{F_{MHz} \times 1.05} \quad \text{feet}$$

where HA is the physical antenna length in feet, and h/λ is the electrical length in fractions of a wavelength. For example, if F_{MHz} = 3.8, then HA = 246 h/λ. Then, if h/λ = 0.25, HA = 61.5 feet.

The results of the first computer program I wrote gave values of R and X vs h/λ. This standard data is reproduced in fig. 1 for use in selecting some examples of practical heights of verticals.

The antenna lengths selected for consideration were even multiples of 10-foot pieces of conduit, and specifically were 30-, 40-, 50-, 60- and 70-feet long. Some special lengths were also selected, and these were lengths which were easily matched to RG-8/U or RG-11/U coaxial cable.

table 1. Calculated vertical antenna characteristics vs height, and vswr performance with three different types of matching systems.

1. No matching network (center of coax connected directly to antenna base).

HA (feet)	h/λ	R	X	coax	vswr
80.0	0.325	90.6	+230.0	RG-11/U	8:1
76.0	0.31	75.0	+180.0	RG-11/U	8:1
60.0	0.243	33.8	+ 5.7	RG-11/U	2.2:1
70.0	0.285	55.0	+108.0	RG-8/U	6:1
69.0	0.28	52.0	+ 95.0	RG-8/U	5:1
61.5	0.25	36.6	+ 21.3	RG-8/U	1.8:1
60.0	0.243	33.8	+ 5.7	RG-8/U	1.5:1
59.3	0.241	32.0	0	RG-8/U	1.6:1

2. Series capacitor, C, between coaxial transmission line and antenna base.

HA (feet)	h/λ	R	X	coax	C,pF	vswr
80.0	0.325	90.6	+230	RG-11/U	182	1.2:1
76.0	0.31	75.0	+180	RG-11/U	233	1.0:1
70.0	0.285	55.0	+108	RG-8/U	388	1.06:1
69.0	0.28	52.0	+95	RG-8/U	441	1.0:1
61.5	0.25	36.6	+21.3	RG-8/U	1966	1.4:1

3. Type-C L-network (see fig. 3) used with RG-8/U coaxial cable. Vswr 1.0:1.

HA (feet)	h/λ	R	X	L, μH	C,pF
60.0	0.2433	33.8	5.7	0.80	591
50.0	0.2028	20.7	-86.5	4.69	991
40.0	0.1622	12.1	-183.8	8.62	1467
30.0	0.1216	6.3	-303.0	13.41	2164

4. Type-A L-network (see fig. 3) used with RG-8/U coaxial cable. Vswr 1.0:1.

HA (feet)	h/λ	R	X	L, μH	C,pF
70.0	0.285	55	+105	4.33	637

Table 1 lists these choices, the network used (if any) and the resulting vswr calculations made at 3.8 MHz. This data shows that with no network and RG-11/U transmission line, the lowest vswr is 2.2:1 with HA = 60 feet. With no network and RG-8/U coax, the lowest vswr is 1.5:1 with HA = 60 feet. If the proper series capacitor is used, the vswr is 1.0:1 for RG-8/U at HA = 69 feet, and for RG-11/U at HA = 76 feet.

For antennas of other heights the vswr is 1.0:1 if the proper L-networks are used. The L-networks were calculated using the methods outlined in my *QST*

article,[4] or from the methods detailed in *ham radio.*[5]

network losses

The next topic to explore is that of matching network losses. Of course, there is no network loss for cases not using a matching network, and I will also assume that there are no losses in any of the network capacitors. However, there are coil losses, and these can be significant for matching networks for short vertical antennas. Assuming that the power output to the matching network is 600 watts, and that the Q of the network coil is 200, coil losses are as shown in **fig. 2.** On the basis of this graph you can select a height of vertical depending upon the amount of network loss that you are willing to accept.

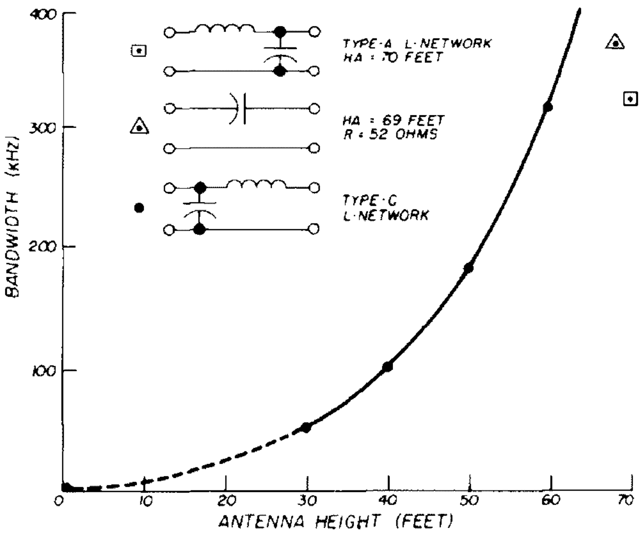


fig. 3. Bandwidth of vertical antennas vs height. Bandwidth edges defined by frequencies at which vswr increases to 2.0:1.

As an example of these calculations, consider a 30-foot vertical, where HA = 30 feet, R = 6.3 ohms, and X_L = 320 ohms. If there are no capacitor losses, the 600 watts delivered by the coax feedline to the network must be divided between the network coil and the antenna. There are earth losses too, but I will consider these in a later article, and assume that they are zero now in order to examine network losses. The coil loss resistance, r, is X_L/Q, or 320/200 = 1.6 ohms. The current in the coil and antenna is then

$$I = \sqrt{\frac{P}{R+r}} = \sqrt{\frac{600}{6.3+1.6}} = 8.71 \text{ amps}$$

Thus, the coil loss is I^2R = 121.5 watts, and I^2R = 478.5 watts delivered to the antenna.

antenna bandwidth

The last topic for this article is that of the bandwidth for the antenna examples given. For each of these examples the network was designed to make the input impedance seen by the coax feedline to be 52 ohms for a vswr of 1.0:1 at an arbitrary frequency of 3.8 MHz. Another computer program was written which calculated the input impedance versus frequency for the same network and same height vertical. The previous program furnished the changing antenna self impedance versus frequency. The changing input impedance versus frequency was plotted on a Smith chart for 30-, 40-, 50- and 60-foot verticals using one type of L-network; the 69-foot vertical using a series C network; and the 70-foot vertical using a different type of L-network. Bandwidth was arbitrarily defined as being the difference between those frequencies having a vswr of 2.0:1. These bandwidths are shown in fig. 3. As expected, the shorter antennas have a smaller bandwidth. Use of this graph will help you to select an antenna height depending upon what bandwidth is acceptable to you.

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1. Jerry Sevick, W2FMI, "The W2FMI Ground-Mounted Short Vertical," *QST*, March, 1973, page 13.
2. Edward C. Jordan, *"Electromagnetic Waves and Radiating Systems,"* Prentice-Hall, New York, 1950, chapter 11 (impedance).
3. John R. Kraus, *Antennas*, McGraw-Hill, New York, 1950.
4. Robert E. Leo, W7LR, "An Impedance-Matching Method," *QST*, December, 1968, page 24.
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ham radio

lowpass filters

for solid-state linear amplifiers

Complete low-power, lowpass filter designs for 160, 80, 40, 20, 15 and 10 meters

G. Kent Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

Semiconductors have finally found their way into high-frequency linear amplifiers. Broadband, untuned amplifiers that were seldom practical with vacuum tubes have become the best approach with semiconductors. Along with the advantages of rapid QSY and circuit simplicity comes the problem of unwanted harmonics. It is a recognized fact that a lowpass filter is a necessity with this type of amplifier.[1] Many of the broadband amplifier designs use push-pull circuitry, which may supress the second harmonic by 50 dB, but the third harmonic is suppressed only 12 dB.[2]

I have selected an elliptic function filter design that provides low insertion loss, low vswr and attenuation peaks at the second and third harmonics. It is assumed that a linear amplifier will be used in the phone portion of the bands but adequate suppression is obtained for CW operation too.

Several listings of normalized filter data have been printed and are quite simple to use.[3,4] Unfortunately, these publications are seldom in the average amateur's library. The theoretical design has been compromised only slightly to allow use of standard mica capacitors (either compression molded or dipped) that can be purchased from Allied, Newark or other suppliers. Five-hundred-volt capacitors will handle several hundred watts if the vswr of the antenna is near unity and are a wise choice unless low power and miniaturization is contemplated. The Micrometals toroidal cores listed are readily available from Amidon Associates.

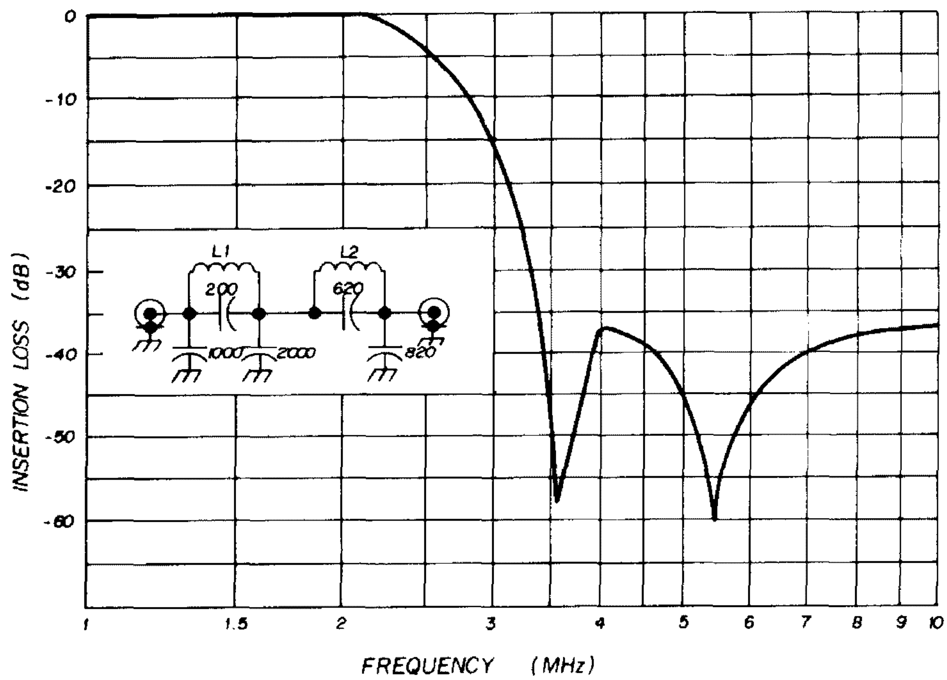

fig. 1. 160-meter lowpass filter. L1 is 26 turns number-18 on Amidon T80-2 toroid (4.2 μH). L2 is 23 turns number-18 on Amidon T80-2 toroid (3.13 μH). Insertion loss is 0.1 dB over the 160-meter band.

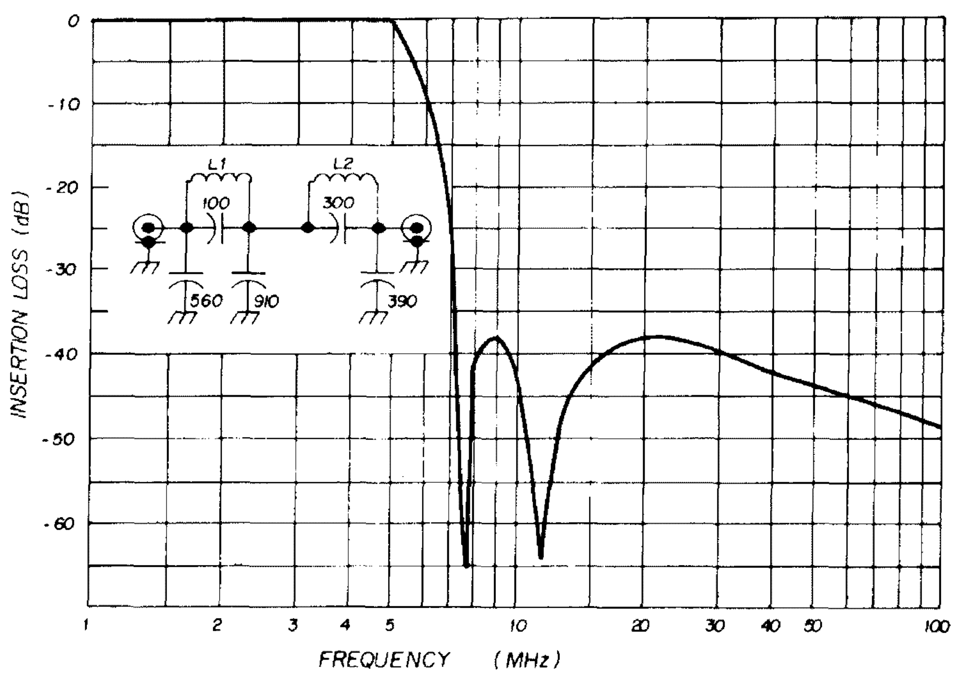

fig. 2. 80-meter lowpass filter. L1 is 18 turns number-16 on an Amidon T80-2 toroid (1.9 μH). L2 is 16 turns number-16 on an Amidon T80-2 toroid (1.46 μH). Insertion loss is 0.12 dB over the 80-meter band.

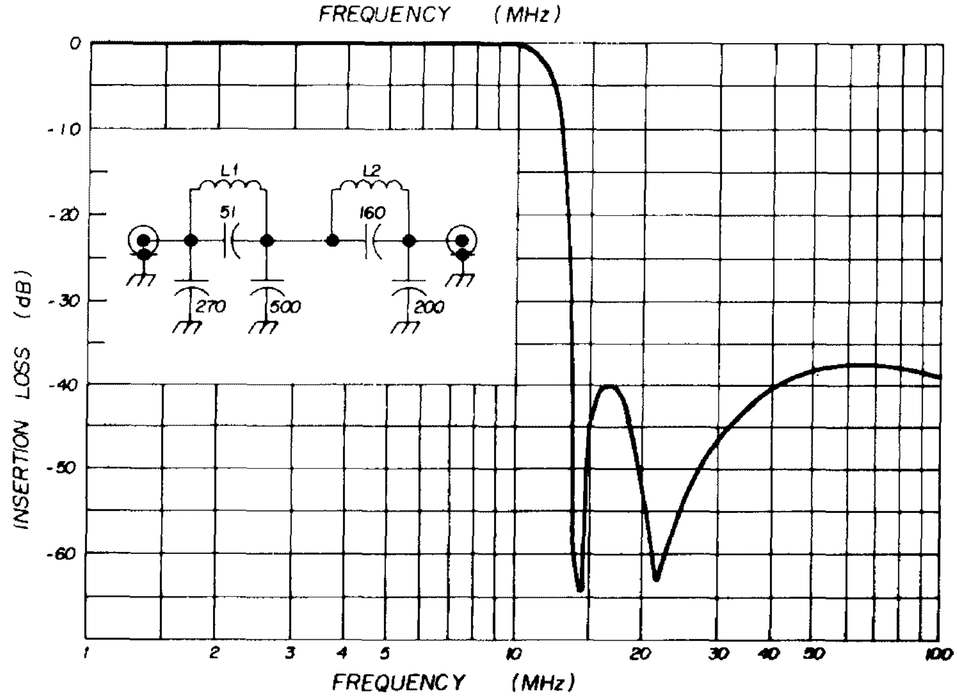

fig. 3. 40-meter lowpass filter. L1 is 10 turns number-16 on an Amidon T80-6 toroid (0.57 μH). L2 is 9 turns number-16 on an Amidon T80-6 toroid (0.41 μH). Insertion loss below 18 MHz is 0.17 dB.

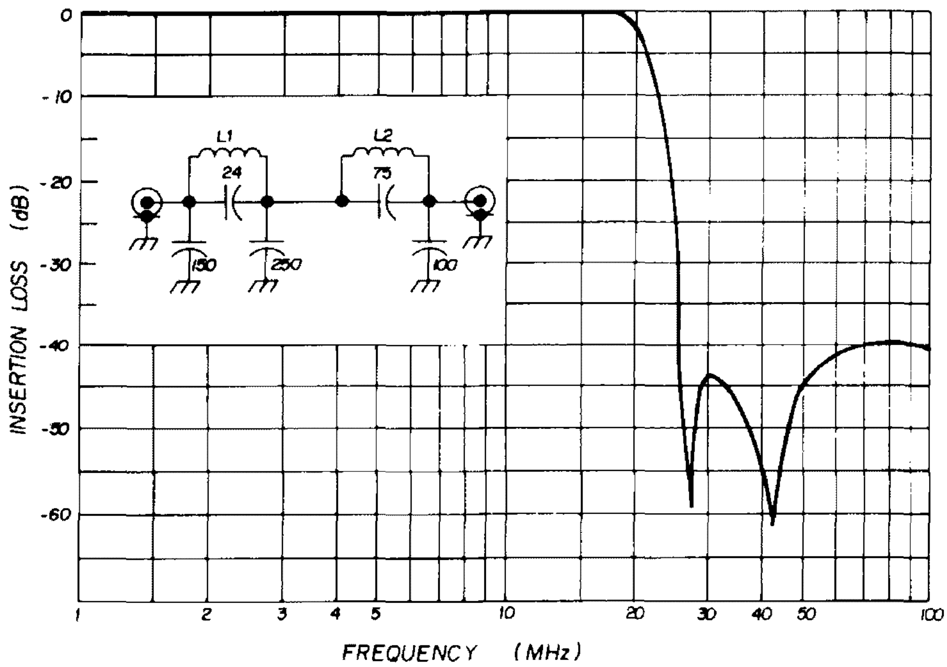

fig. 4. 20-meter lowpass filter. L1 is 10 turns number-16 on an Amidon T80-6 toroid (0.57 μH). L2 is 9 turns number-16 on an Amidon T80-6 toroid (0.41 μH). insertion loss below 18 MHz is 0.17 dB.

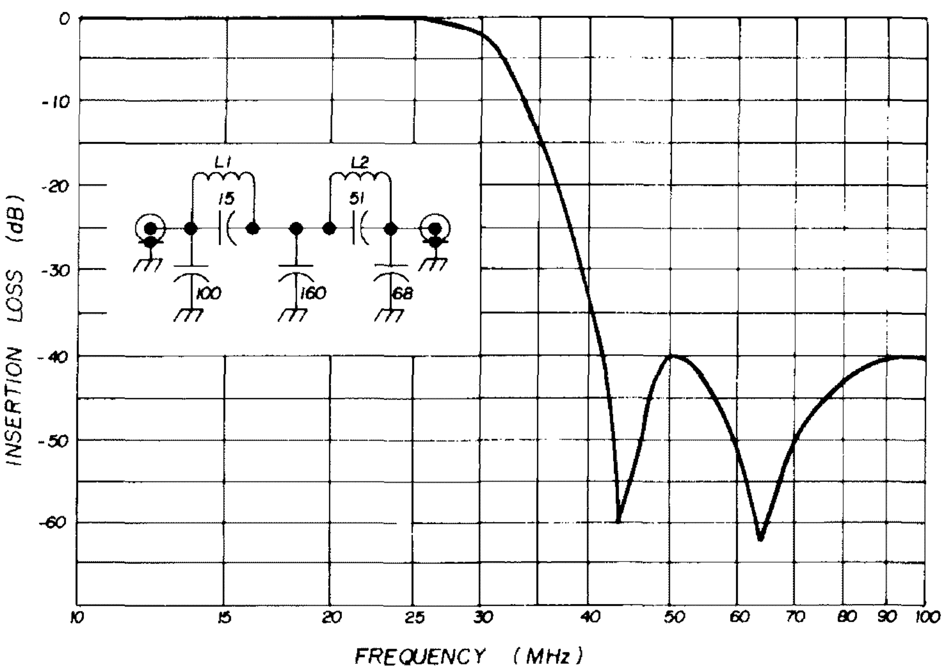

fig. 5. 15-meter lowpass filter. L1 is 9 turns number-16 on an Amidon T80-6 toroid (0.41 μH). L2 is 8 turns number-16 on an Amidon T80-6 toroid (0.27 μH). Insertion loss below 24 MHz is 0.25 dB.

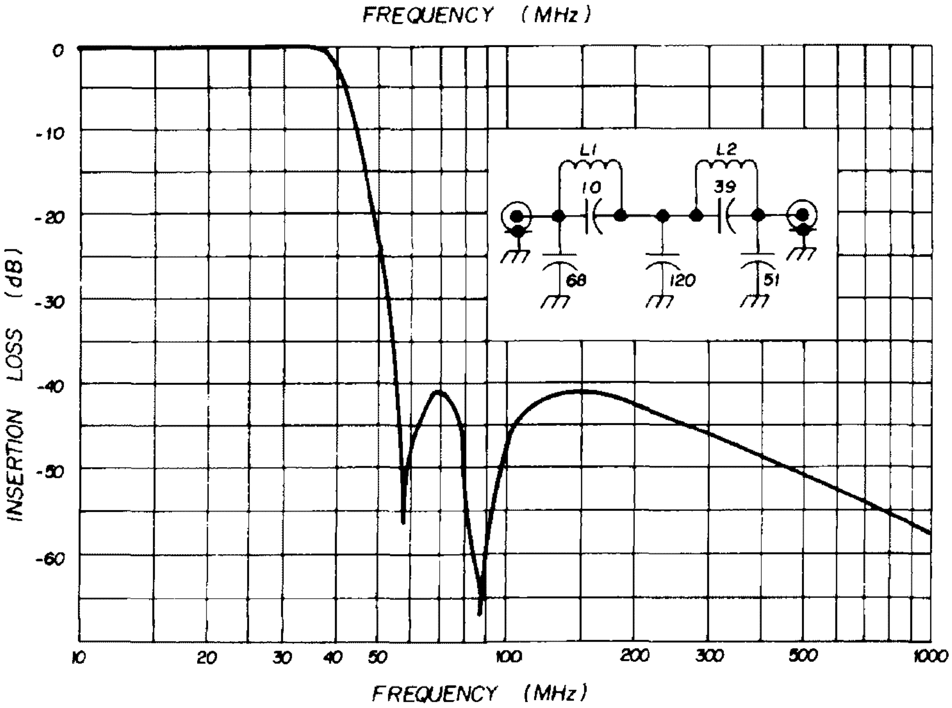

fig. 6. 10-meter lowpass filter. L1 is 8 turns number-16 on an Amidon T80-6 toroid (0.33 μH). L2 is 7 turns number-16 on an Amidon T80-6 toroid (0.19 μH). Insertion loss below 35 MHz is 0.3 dB.

Bathtub cans from old oil-filled capacitors make good cases with BNC or RCA Phono connectors soldered in the ends. If you are going solid state those old 1000-volt, oil-filled capacitors won't be needed anymore, so salvage the cases.

Designs and winding information are given for the top six amateur bands. The response curves were obtained by computer analysis with actual testbench verification of the 80- and 40-meter filters (those are the two bands I'm now working). The filters perform well without tuning, but a little adjustment of the resonant frequency will help assure 60-dB suppression of the second and third harmonics. There are several ways to tune the toroidal resonators.[5] In all six filters inductor L1 will resonate at the third harmonic and L2 will be resonated at the second harmonic.

Please join me in fighting air pollution. Keep the bands upstairs clean for the other operators!

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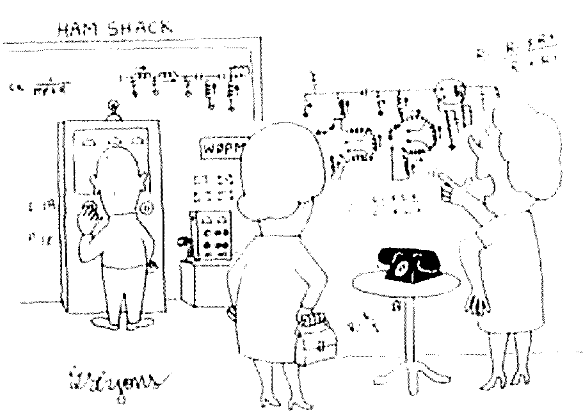
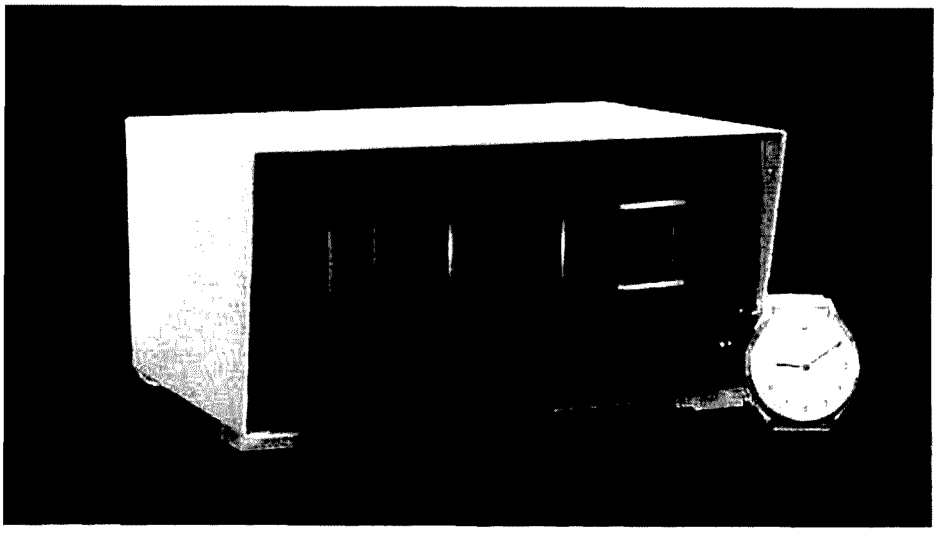
4. Anatol I. Zverev, *Handbook of Filter Synthesis,* John Wiley & Sons, New York, 1970.
5. G. Kent Shubert, WAØJYK, "Tuning Toroidal Inductors," *ham radio,* April, 1972, page 24.

ham radio



"That's Dave's bad habit!"



a simplified digital readout system

Discussion of a new and unique system of glyphs which are ideally suited for use with IC logic

Harold W. Thompson, W6OIS, President, Rad-Ex Syntactics*

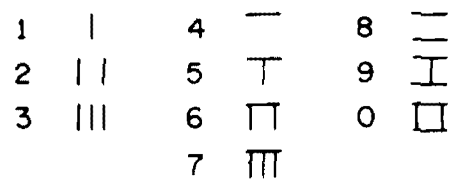
The word instrumentation has a magic ring and is something every true amateur desires, but usually cost tempers that desire. Some years ago, while fighting to read a Potter counter which used four columns of four lamps each, displaying a BCD format, I was sure there must be a better way. As transistors mutated to integrated circuits and costs plummeted, a number glyph was developed that could revolutionize numeral concepts. As not everyone is interested in the historical development of numerals, suffice it to say that man has spent several thousand years developing and changing his numeric glyphs. Now the age of the computer will require another change if man and machine are ever to communicate directly.

invention

Accepting the fact that digital equipment will never be anything more than

*Rad-Ex Syntactics, 1043 East Atchison Street, Pasadena, California 91104

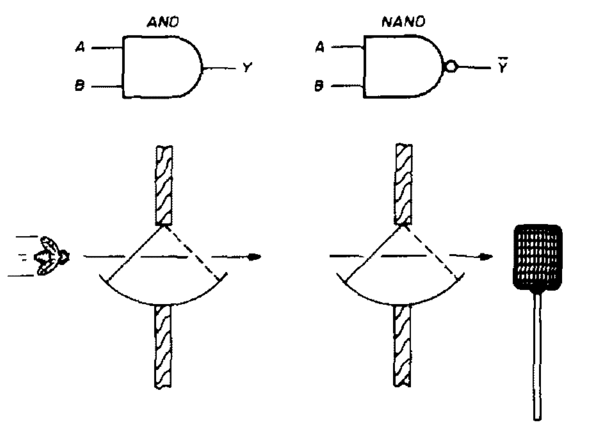
off and on switches relegates them to *binary* operation. Man must learn to use these bits of information in an efficient manner. My own approach to this binary age is to reform the old Potter readout lamps by arranging them in illuminated bars as follows:



Why? These bars are a *direct* output from the decade counter. What has been gained? A readily readable symbol, reduced cost, simplified circuitry and more reliability.*

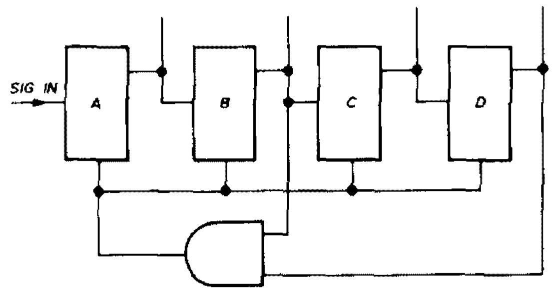
logic

To answer in more detail, some discussion of logic, counters and circuitry is necessary. It helps to remember that any digital device is just a mass of electrical switches. It depends on your own insight and knowledge to parallel or series them



to produce your desired result. For instance, the door and screen door are AND gates. Both must be open for the fly to come in. The results can be negative, NAND, if somebody is waiting with the fly swatter. The same analogy can be applied to a door and window forming the OR and NOR gates.

Master these concepts and their ramifications, and you will be able to follow any logic design. One ingredient you must apply to all logic is that in the real world a finite time must be allowed for signals



D	C	B	A	
0	0	0	0	0
0	0	0	1	1
0	0	1	0	2
0	0	1	1	3
0	1	0	0	4
0	1	0	1	5
0	1	1	0	6
0	1	1	1	7
1	0	0	0	8
1	0	0	1	9
1	0	1	0	10 = Reset signal
0	0	0	0	

fig. 1. Theoretical divide-by-10 ripple counter.

to pass through each logic block. Believe me, this is important. Usually, logic circuits end up in a counter, probably for display, or register, a group of binary bits which will be processed further.

The device of most interest to the amateur is the ripple counter. **Fig. 1** is the logic diagram for a ripple counter decade unit. This unit should go through the binary sequence shown in the accompanying table and repeat.

The old problems of finite time and reality interfere and develop into a condition called *critical race.* Beware of these problems. It's been the downfall of many designs.

What happens is that the reset pulse has set all flip-flops to zero but in so doing, flip-flop B passes a trigger signal to flip-flop C and you find yourself with binary 0100. Thus, you spend a few more hours devising lockout circuitry so this

*U.S. Patent No. 3,671,943. A copy may be obtained from the Commissioner of Patents, Washington, D.C. 20231 (50 cents money order).

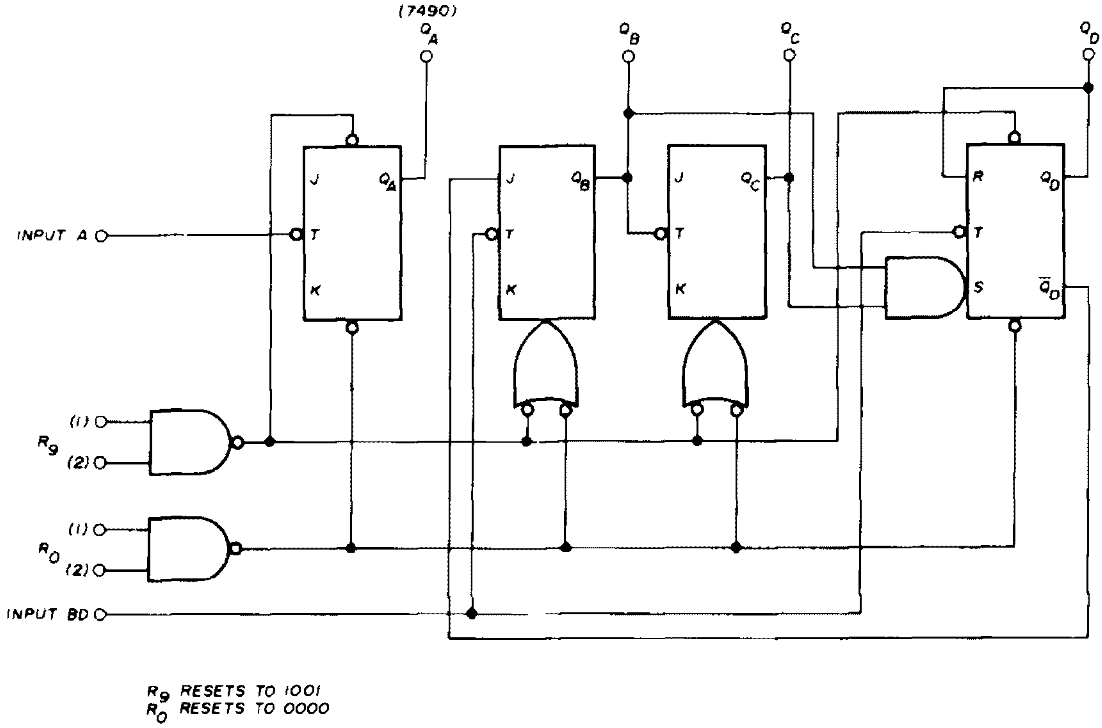


fig. 2. Commercial decade counter.

can't happen (see fig. 2, a commercial decade counter). Now that your counter is designed, you will need an indicator to communicate the binary number it holds.

About 30 years ago Potter built one of the first counters which, with four neon lamps per decade, displayed the state of each flip-flop. Each lamp was weighted 1, 2, 4 or 8 and you simply added the values, or, to be more precise, you learned binary equivalents for Hindu-Arabic numbers.

Engineers soon mastered the art of using gates. From then on it didn't take too much time to build decoding logic which could combine the counter outputs and light individual one-of-ten lamps such as the Nixie tube. Finally, the ten outputs were combined into a diode matrix to illuminate a 7-segment numeral (see fig. 3).

Rad-Ex began a search for a readout display which could be connected directly to a counter and would convey a feeling of quantity. The Rad-Ex numerals fit this concept perfectly and in surprising ways open up several interesting possibilities such as handwritten machine readable characters, and a new way to teach arithmetic.

technical

Electronically, the basic counter is different and more simple than present decade counters, has no critical race

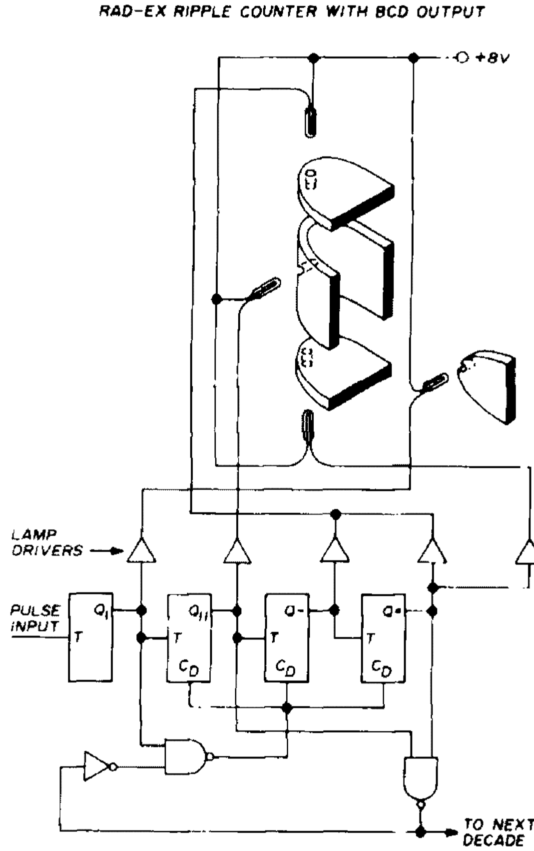


fig. 4. Rad-Ex System, decade counter to readout.

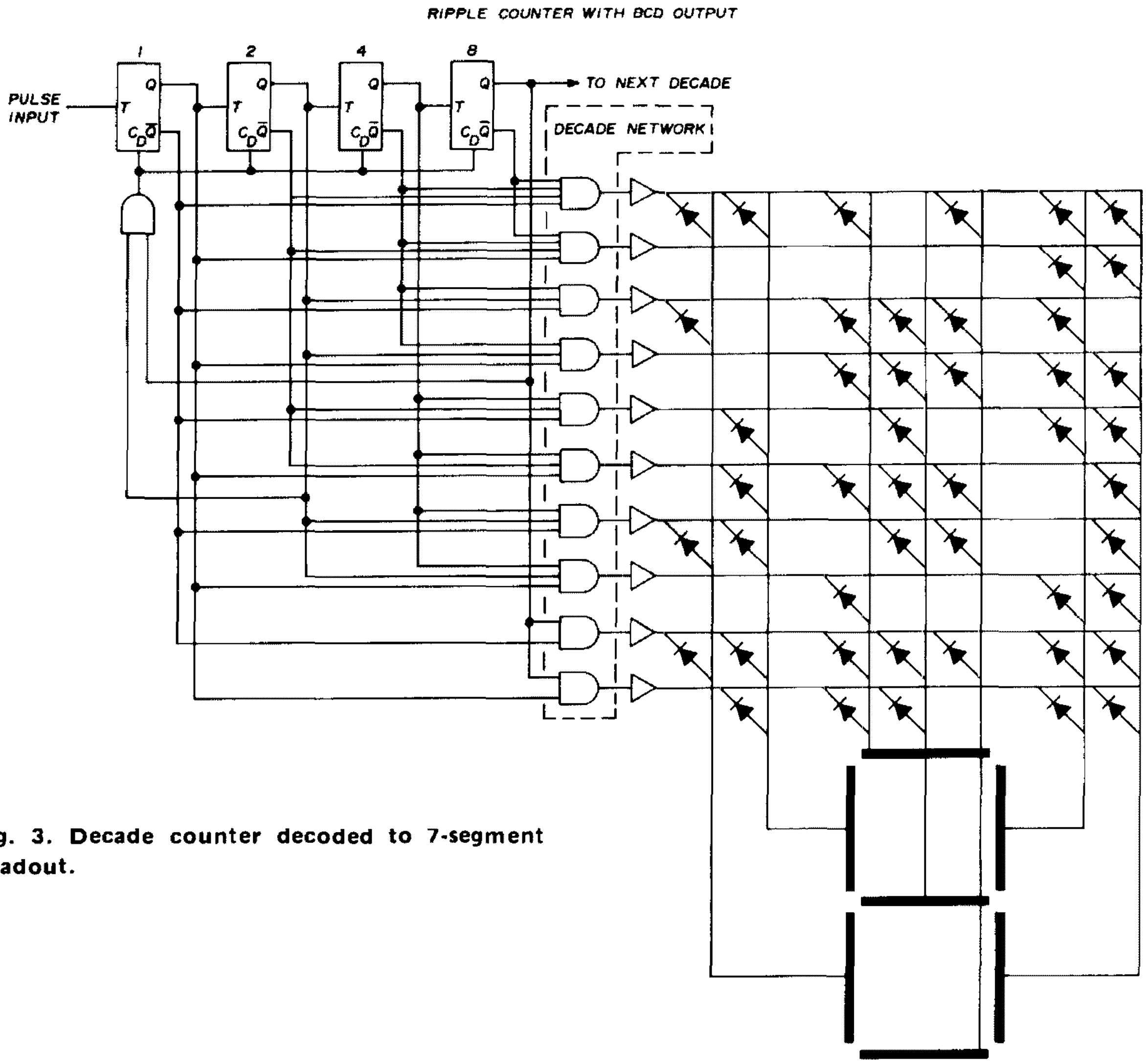


fig. 3. Decade counter decoded to 7-segment readout.

paths, while almost doubling the counting rate and automatically suppressing leading zeros. **Fig. 4** shows the circuit; its truth table is below:

D	C	B	A	
0	0	0	0	State Before Count Begins
0	0	0	1	
0	0	1	0	
0	0	1	1	
0	1	0	0	
0	1	0	1	
0	1	1	0	
0	1	1	1	
1	0	0	0	
1	0	0	1	
1	0	1	0	
1	0	1	1	AND to Reset
0	0	0	1	

At first the truth table seems to indicate two zeros, but the fact is that binary 0000 is a true starting point where no count exists. It is also a unique way of suppressing leading zeros. Once the counter starts, binary 1010 becomes the systems designated zero. (For many reasons it is handy to have a binary designation for zero; for instance, on a punch tape a blank space is ambiguous.) Electronically, good things happen too. Circuitwise, the first flip-flop is independent and needs no clear signal when counting. Thus critical races are eliminated, allowing the use of a simple 3-input NAND gate to reset the counter for each decade count.

For readout purposes, with one exception, the lamps are tied through lamp drivers directly to the flip-flops. The one exception is the case where binary 1000 also illuminates the lamp associated with flip-flop C.

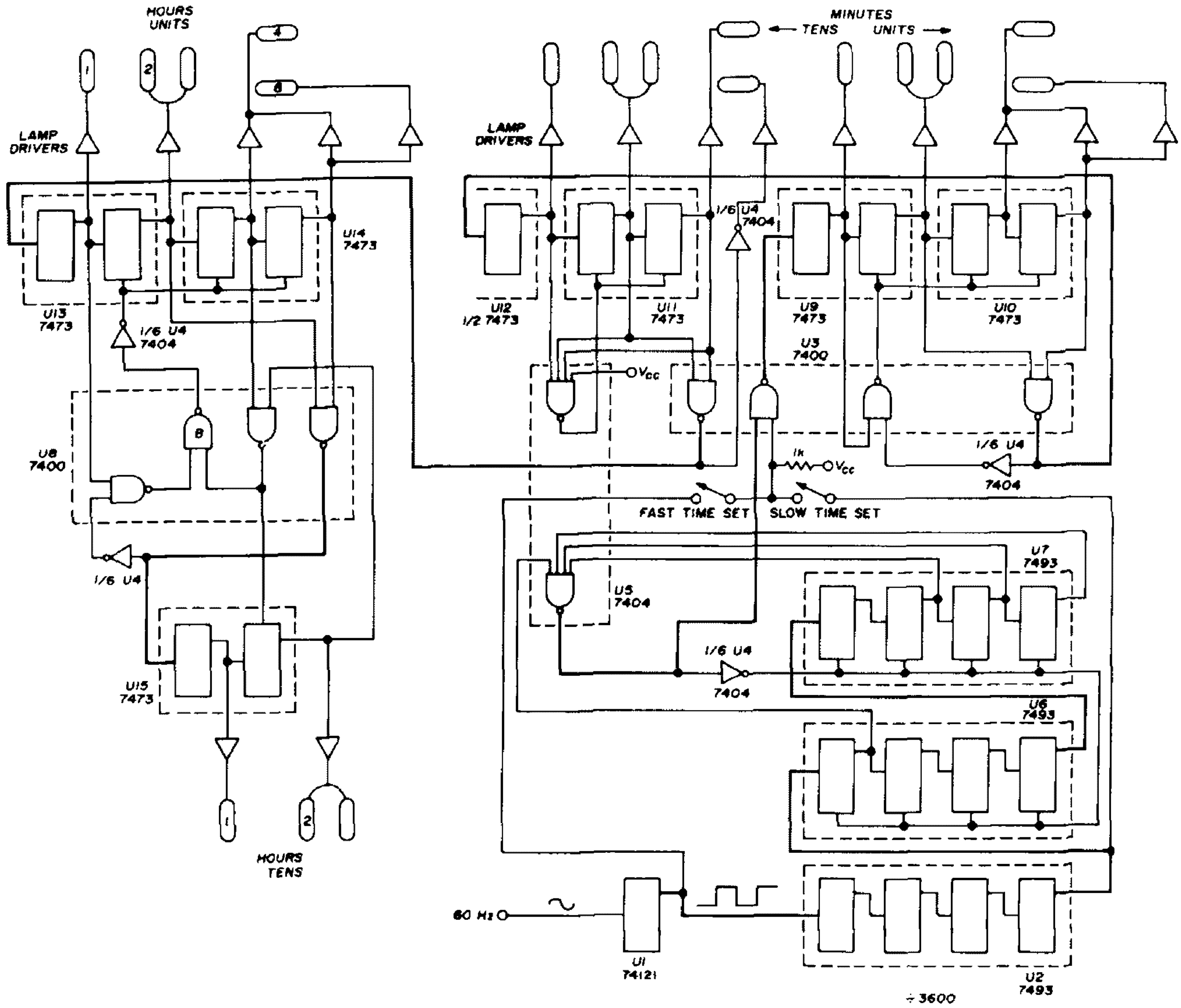


fig. 5. Twenty-four hour clock using Rad-Ex readouts.

application

The use of the Rad-Ex system can best be illustrated in a 24-hour clock where mixed counting is necessary. Fig. 5 is the logic diagram for the clock and should be used as reference in the following sequence of operation: 60 Hz is fed to a one-shot multivibrator, U1, simply to eliminate false triggering. The output from U1 triggers a chain of 7493 flip-flops U2, U6, U7, which are used as a ripple counter to divide incoming pulses by 3600 (1110, 0001, 0000).

All that is required for this operation is to detect the four binary ones and use a 4-input gate, U5, to reset U6 and U7 (U2 is already at 0000). The resulting one-minute pulses are fed to the first decade counter which displays zero to 9 minutes. As the transition from binary 1001 to 1010 occurs, a pulse passes to a divide-by-6 counter. Again, the non-conventional system works to the advantage of simplicity. The truth table shows how the Rad-Ex numbers zero through 5 are energized.

C	B	A	
0	0	1	
0	1	0	
0	1	1	
1	0	0	
1	0	1	
1	1	0	
1	1	1	AND to Resets B & C
0	0	1	

The 24-hour portion of the clock uses the standard Rad-Ex decade counter plus a counter to keep track of the hours.

Essentially, the decade counter acts the same as the minute decade with the exception that the suppressed leading zero feature may be observed. The binary numbers 10 (2) and 0100 (4) are combined in a NAND gate, U8, to reset both decade and modulo-4 counters to binary

A check of the cost of ICs for this unit amounts to $9.85. ICs for a comparable clock designed to use 7-segment readouts cost about $14.00. A set of Rad-Ex readouts costs $4.75; the 7-segment readouts and four decode units cost about $15.00.

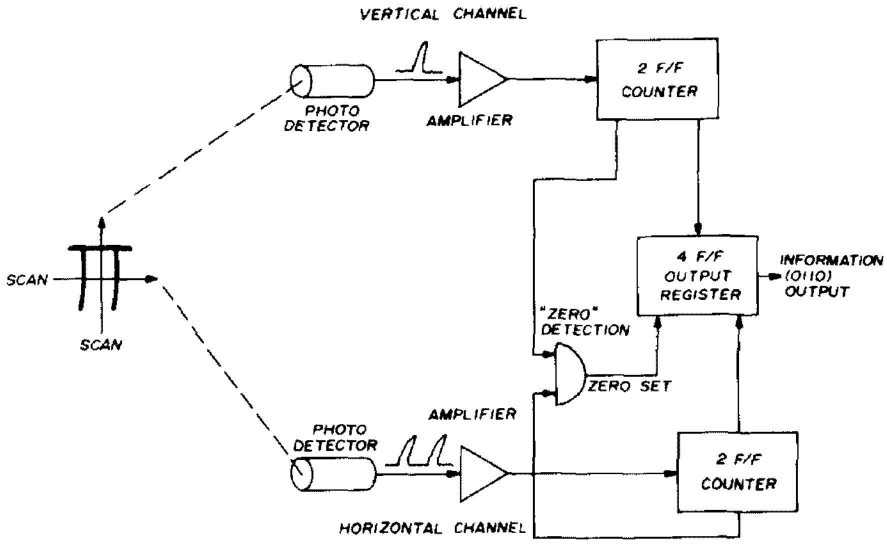
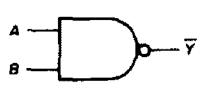


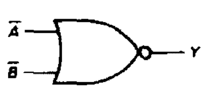
fig. 6. Rad-Ex optical reader device.

(00) (0000). Thus, until the first hour is reached only the minute display is illuminated.

The sharp eye may spot two NAND gates, part of U3A and U8B, which don't seem quite right. Here you must juggle your logic thinking for in the case of the NAND gate, I say, "If both inputs A and B (A•B) are high, Y is low ($\overline{Y}$)."



You may look at this in a different way and say, "If A or B (A+B) is low, ($\overline{A+B}$), Y is high.



These two NAND gates are used in a NOR sense, U8B is used to clear the decade counter whenever the count of ten or 24 is present. Gate U3A is used to allow a fast or semi-fast setting of the clock.

future

As mentioned earlier, this is a symbol that can be handwritten and machine read. Fig. 6 is a block diagram showing a method of scanning the numeral to set up a BCD character. The number is scanned horizontally to pick up ones and vertically to pick up fours. These outputs are combined in a shift register which on command transmits the BCD word.

Hopefully, this article will inspire experimenters, hobbyests, and professionals to become involved in the man-machine communication problem. At present, there is a growing need for man to communicate through handwritten glyphs directly to computers, with bank drafts and zip codes being prime examples. There is also need to simplify instrumentation so that digital voltmeters, frequency counters, etc. can talk more directly and less expensively to you. Rad-Ex Syntactics feels we have brought these goals closer to realization.

ham radio

new fets

simplify bias problems

The operating characteristics of several new field-effect transistor families simplify proper biasing

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

Many circuits in the amateur literature specify the Motorola MPF102 fet. The Texas Instruments 2N3819 and the Siliconix U183 have essentially identical data sheets to that of the MPF102. The trouble with all three of these devices is the 10-to-1 spread of I_{DSS} (2-20 mA) which makes bias point and performance somewhat unpredictable. I_{DSS} is the drain current when zero bias is applied between the gate and source terminals of the fet.

bias problem

The circuit shown in **fig. 1** is a typical fet biasing arrangement. The gate of the fet is held at ground potential (zero volts) by resistor, R_g, and the voltage drop across the source resistor, R_s, biases the source terminal to some voltage above ground. Thus, the gate is biased negative with respect to the source. The value of the gate-to-source bias is equal to the product of the drain current, I_D, in mA, and the source resistor, R_s, in kilohms. If R_s is 2000 ohms and I_D is 1.5 mA, the voltage across R_s is 3 volts. The gate is thus biased 3 volts more negative than the source.

Fig. 2 shows how drain current varies versus gate-to-source voltage for an fet whose I_{DSS} is 20 mA. This fet could be an MPF102, a 2N3819 or a U183. A straight line is drawn through the origin which represents a source resistor, R_s, having a resistance of 1000 ohms. Notice that a change of one volt along this line

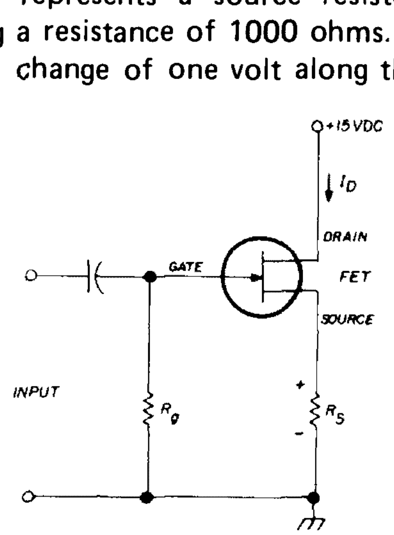


fig. 1. Typical field-effect transistor biasing arrangement.

results in a change in current of 1 mA, indicating a resistance of 1000 ohms.

The point where the straight line intersects the curve gives the values of drain current and gate-to-source voltage. In this case the drain current is about 3.7 mA, and the gate-to-source voltage (drop across R_S) is about 3.7 volts. **Fig. 3** shows how the situation is changed if the fet is replaced by one having a value of I_{DSS} equal to 2 mA. The drain current is now 0.48 mA, and the gate-to-source voltage is 0.48 volt. This shows that the drain current of an MPF102, 2N3819 or U183 may be anywhere from 0.48 mA to 3.7 mA when a 1000-ohm source bias resistor is used.

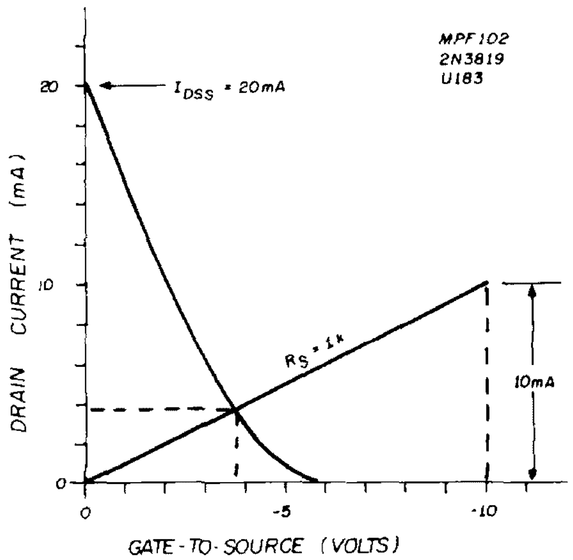


fig. 2. Typical fet drain current vs gate bias characteristic for I_{DSS} = 20 mA.

With such a wide range of possible drain current, it is impossible to choose an efficient drain load that would be suitable for all fets of these types. As an example, suppose a 2N3819 is to be used in a resistance-coupled audio amplifier stage such as shown in fig. 4. Under no-signal conditions, it is desired that the dc drain voltage be 10 volts. This means there must be a 5-volt drop across the drain resistor, R_D. If a 2N3819 is used which has an I_{DSS} of 2 mA, the drain current will be 0.48 mA, and the value of R_D should be

$$R_D = \frac{5 \text{ volts}}{0.48 \text{ mA}} = 10.4\text{k ohms}$$

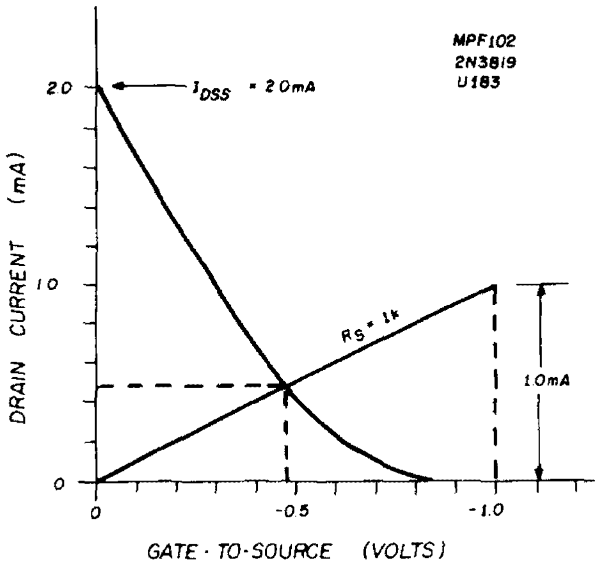


fig. 3. Typical fet drain current vs gate bias characteristic for I_{DSS} = 2 mA.

But if a 10k resistor is used for R_D, and a 2N3819 with I_{DSS} equal to 20 mA is plugged into the circuit, the drain current will *try* to be 3.7 mA, which would produce a 37 volt drop across R_D. Obviously this is impossible with a 15-volt supply, so the fet simply saturates, and linear amplification is not possible. If, on the other hand, R_D is chosen so that it has a 5-volt drop when the current through it is 3.7 mA (I_{DSS} equal to 20 mA), its value would be

$$R_D = \frac{5 \text{ volts}}{3.7 \text{ mA}} = 1.35\text{k ohms}$$

If this value of resistor is used with an fet

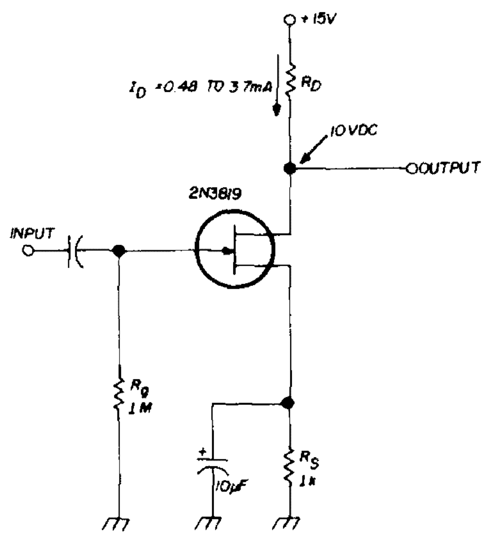


fig. 4. Simple audio amplifier circuit using a 2N3819 fet.

having an I_{DSS} of 2 mA, the drop across it will be only

(0.48 mA) (1.35k) = 0.65 volt

The problems involved in biasing fets with large I_{DSS} spreads should now be quite apparent.

new fets give relief

One way to get around this biasing problem is to take a large number of fets and sort them into groups, each group having a relatively narrow I_{DSS} range. Fortunately, manufacturers are now doing this. Texas Instruments has taken the 2N3819 and broken it into five fet types, each of which has an I_{DSS} spread of 2 to 1 or less. These fets, which have a different pin configuration than the 2N3819 are listed in table 1. All of these

table 1. List of 2N3819-type fets with small I_{DSS} spreads.

fet type	I_{DSS}
2N5949	12-18 mA
2N5950	10-15 mA
2N5951	7-13 mA
2N5952	4-8 mA
2N5953	2.5-5 mA

fets are priced under a dollar in small quantities, and they should be available from any of the larger electronic wholesalers which stock TI semiconductors.

Fig. 5 shows how the drain current of a 2N5953 would be in the range of 0.7 to 1.1 mA if its source bias resistor is 1000

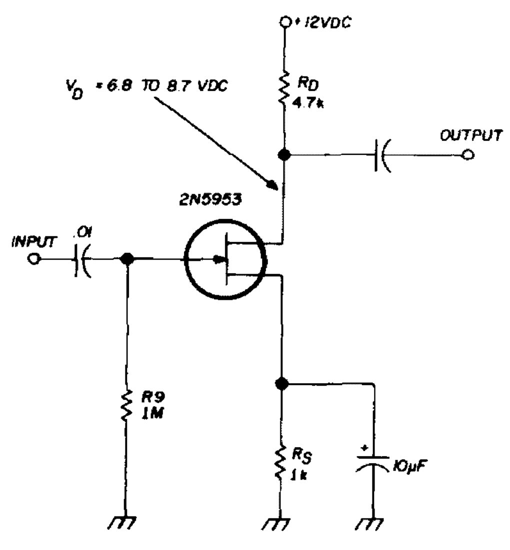


fig. 6. Audio amplifier circuit using a 2N5953 fet.

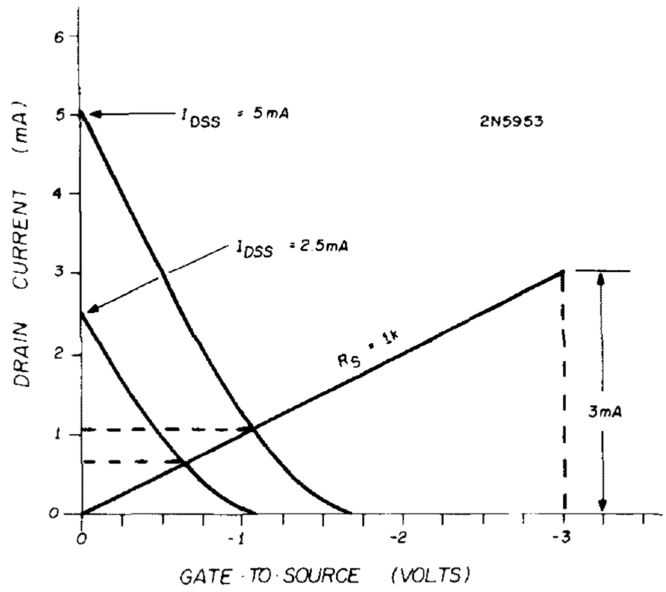


fig. 5. Typical drain current vs gate bias characteristic for Texas Instruments 2N5953 fet.

ohms. The drain current could be said to be 0.9 mA, ± 0.2 mA for all 2N5953 fets used with a 1000-ohm source bias resistor. **Fig. 6** shows a practical fet audio amplifier circuit using the 2N5953. Voltage gain is typically around 10, and any 2N5953 used in this circuit will be reasonably well biased.

The Motorola 2N5484 series, priced at about a dollar each, have I_{DSS} spreads of 5 to 1 and 2.5 to 1.

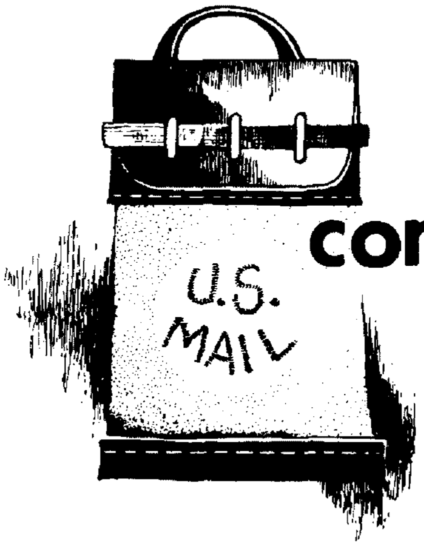
fet type	I_{DSS}
2N5484	1-5 mA
2N5485	4-10 mA
2N5486	8-20 mA

This isn't as good as the TI 2N5949 series, but is considerably better than the MPF102 types.

conclusion

The newer fets, having lower I_{DSS} spreads, allow the use of simple bias arrangements to arrive at reproducible circuits. The cost of these devices is not much higher than the older types having wide I_{DSS} spreads. Thus, fets may be applied with greater ease to a wide variety of circuit applications, and the high input impedance of fets may be taken advantage of without the penalty of unpredictable bias conditions.

ham radio



comments

log-periodic antennas

Dear HR:

Last December the Cyprus Government issued me a ham ticket. Then I began a mad scramble, looking for a source of commercial beam antennas. I quickly found that all antennas cost at least double the retail prices in the United States. Considering the cost of a beam, tower and rotator, it was very discouraging. Next, I searched for material for building my own beam. Aluminum is practically unobtainable here, and PVC tubing was out because the available material is too thin and flexible.

The log-periodic antenna article by W4AEO in the September, 1972, issue of *ham radio* provided the answer. It was the only antenna I could find parts for and the design had sufficient gain to consider working the United States (as noted by myself and others, most of the signals from the States on 20 meters average 20-dB less here than reported by stations located in continental Europe).

Putting the log periodic together was much faster than I imagined, but a vswr of 4:1 when fed with 50-ohm coax was a puzzle until a check with a noise bridge indicated the input impedance was 200 ohms. I wound a 4:1 balun and maximum vswr on 15 and 20 meters is now 1.2:1. On-the-air reports indicate at least 8-dB gain. I'm running a Collins KWM-2 barefoot, but many old timers accuse me of using at least a kilowatt.

It goes without saying that W4AEO's log periodic, while requiring some acreage, provides considerable gain and solves the parts and money problem. Due to W4AEO's article, a number of us in Nicosia are building other log periodics. We are discovering that the surface has only been barely scratched, and amateurs still have plenty of elbow room to incorporate their own innovations.

Stan Whiteman, 5B4AO/W1MDZ
Nicosia, Cyprus

reciprocating detector

Dear HR:

I have received several letters regarding my "reciprocating detector" article which appeared in the March, 1972, issue of *ham radio*. Transistor Q5 is part of the reciprocating detector switch, but the questions are understandable due to the lack of a dot to show a connection in the schematic; resistors R4 and R5 should be joined with a dot where these two resistors form a junction point at the input to the diode and the base of Q5. The diode is a 1N252.

Several readers have also asked where the selectivity curve is 500-Hz wide and what is its slope. The filter I used was designed to have its 500-kHz passband at the 3-dB points on a slope which is not particularly steep for an inductive filter. Indeed, at 500 Hz, the L3 inductance is very loosely coupled to the other two sections of the transformer. The bandpass formula (f_r/Q_o) indicates that the bandpass of the filter is actually narrower than

500 Hz – in fact, bandpass is closer to 250 Hz. The 390-ohm resistor used in series with one of the differential inputs loads the thing down so it is broader. If the bandpass is too narrow, poor lock-in range is experienced on a-m, and there is very poor "presence" in the quality of ssb signals. If the bandpass is too wide, poor impulse rejection will result.

Stirling M. Olberg, W1SNN
Waltham, Massachusetts

vhf fm in the United Kingdom

Dear HR:

Fm channel operation in the United Kingdom is now going strong, thanks to the imported black boxes and new regulations permitting 12.5-kHz deviation (there are lots of 25-kHz mobile equipment around, made by Pye and Storno). We now have one repeater working north of London in Hertferdshire. The callsign is GB3PI with input at 145.15 MHz and output at 145.75 MHz (600-kHz spacing). This repeater just covers outer N. London.

Our Radio Society of Great Britain now has at least five applications for repeaters, and our group, the UK FM Group (Southern), hopes to be able to put a repeater in Hampshire (one of five). Coverage of this repeater should be from Southampton to the edge of southwest London.

John Akam, G8BIH
Wooteys, Alton, Hants

finding square roots

Dear HR:

The technique for computing square roots described in the *ham notebook* by K9DHD in the September, 1973, issue can be extended to increased accuracy by carrying out further iterations. For example, if one uses the first approximation of the square root of 54 obtained as the next estimate and recomputes, an answer of 7.348469 is obtained; this closely approximates the 7.348692 provided by the square root key on my calculator.

As another example, my calculator gives the square root of 75 as 8.660254. Using the Mechanic's Rule, first *iteration* results in an answer of 8.6875; the second then becomes 8.6602965, and the third the desired 8.660254 whose square is 74.999999. These answers were obtained using the very crude first estimate of 8.

Fred R. Scalf, Jr., K4EID
Springfield, Virginia

Dear HR:

I would like to add a note to the short article on finding square roots which appeared in the September, 1973, issue of *ham radio*.

In that article K9DHD gave a procedure for estimating the square root of any number. By a simple extension, arbitrary roots of any number can be determined with a little work and a hand-held digital calculator. To find the nth root of any number P, estimate the root, X_o, and use the following formula

$$X_1 = \frac{1}{n}\left[(n-1)\,X_o + \frac{P}{X_o^{n-1}}\right]$$

where X_1 is the desired root. For the case of a square root, n = 2, this formula is the same as that presented by K9DHD. For a cube root, n = 3, and the formula is

$$X_1 = \frac{1}{3}(2X_o + \frac{P}{X_o^2})$$

For example, to find the cube root of 30, first estimate the cube root (about 3.1). Then

$$X_1 = \frac{1}{3}\left[(2 \times 3.1) + (\frac{30}{3.1^2})\right] = 3.107$$

This is very close to the accepted, approximate, cube root of 30, 3.1072. Of course, the closer your initial estimate, the closer the answer will be to the exact value. With a little patience, this formula will do great service to anyone making use of it.

Stephen R. Alpert, W1GGN
Auburn, Massachusetts

the ham notebook

surplus thumbwheel switch modification

At this year's Rochester Hamfest I picked up at a bargain price an assembly of thumbwheel switches made for Fairchild consisting of 20 Digiswitch C units. I needed these to complete my frequency synthesizer.[1] I took a chance on these switches in spite of the fact that the connector terminals plainly showed they were coded 1 2 4 2' (the last number is read, "two prime") and not the 1 2 4 8 BCD called for in most synthesizers. I took this calculated risk since, first, I am a cheapskate and the price was too good to resist, and second, I hoped to convert the switching to the required coding.

I am happy to say that I was successful in converting the switches to the desired coding, and I offer the following for other bargain hunters since these switches appear to be in plentiful supply and will probably be showing up on the surplus market. Since the old 1242' code is passé now, these switches can be purchased for one-tenth of their original cost. The time required to rework each switch amounts to 7 to 10 minutes, so one evening's work can yield all the switches necessary for a two-frequency synthesizer.

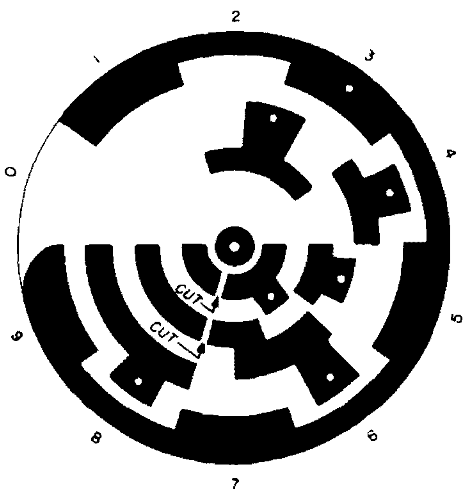


fig. 1. First step in the modification is removal of two thin sections of conducting material on the wiper side of the switch.

These switches can be identified by placing a single unit so the thumbwheel number is right side up and facing to the right. The edge connector coming out the back then indicates, from bottom to top, 335-1, which I presume is the model number, C (common), and 1 2 4 2'. To modify the switch, remove the PC board from the case and set it on the bench so the inside is facing upwards, and the gap is oriented at 9 or 10 o'clock. Using an Xacto knife, remove a thin sliver of the conducting material from the second and third contacts from the center as shown in **fig. 1**. The cut edges should be bevelled so the moving contact can slide up and over easily.

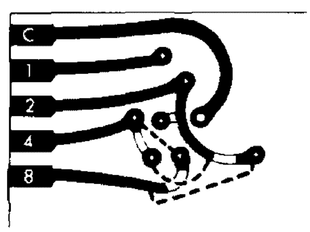


fig. 2. On the reverse side of the switch remove the sections indicated and install the three jumpers.

At the same time, remove the 2' designation and scratch in an 8 in its place. Turn the PC board over and, referring to **fig. 2**, cut away all the conducting material indicated by the crosshatch. From a small length of approximately no. 22 stranded wire use a single strand to solder in the three jump-

ers where shown. Be sure the solder doesn't lump up since these switches have minimal clearance between units when stacked. For best results use a small soldering *iron*.

table 1. With an ohmmeter connected between the common and pins 1, 2, 4 and 8, respectively, of the modified Digiswitch, you should obtain the following readings.

number	1	2	4	8
0	open	open	open	open
1	short	open	open	open
2	open	short	open	open
3	short	short	open	open
4	open	open	short	open
5	short	open	short	open
6	open	short	short	open
7	short	short	short	open
8	open	open	open	short
9	short	open	open	short

When you are finished, put the two parts together and check out the switching sequence to make sure it agrees with table 1. Use an ohmmeter, one lead on C and the other lead on 1 2 4 and 8, respectively, to make sure the BCD sequence is correct.

Geo. Hrischenko, VE3DGX

reference

1. A.D. Helfrick, K2BLA, "High-Frequency Frequency Synthesizer," *ham radio,* October, 1972, page 16.

cutting a minibox down to size

Although several sizes of metal boxes and chassis are available to experimenters, sometimes the nearest size for your project is a little too large. It is not difficult to reduce the *height* of a two-piece metal box. One-half of the box is the top and two ends; the other half is the bottom and two sides.

From the first piece cut down the *ends* to the desired height. On the other piece cut down the *sides* to the same height. The result is a Minibox that fits and looks as good as the original, but which has less height.

I. Queen, W2OUX

finding the focal length of surplus microwave dish antennas

The focal length of most parabolic dish antennas can be determined with two simple measurements, the diameter and the depth as shown in **fig. 3**. The antenna's surface can be described by

$$Y^2 = 4Px$$

where P represents the distance from the vertex to the focal point. The equation is that of a parabola with its vertex at x = 0, Y = 0. The curve is symmetric about the x-axis and opens to the plus-x direction. The coordinates of one point other than the vertex are needed to determine the

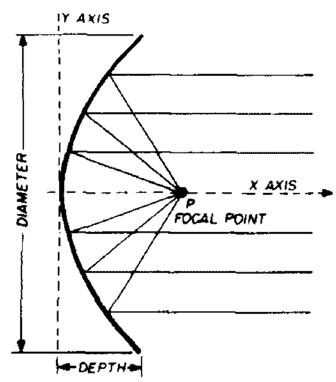


fig. 3. Cross-section of a typical microwave dish antenna. Equation for finding the focal point is given in the text.

curve. The edge of the antenna is a convenient point. The diameter is equal to 2Y and the depth is equal to x. Solving for the focal length P

$$P = \frac{Y^2}{4x} = \frac{\left(\frac{\text{diameter}}{2}\right)^2}{4\,(\text{depth})}$$

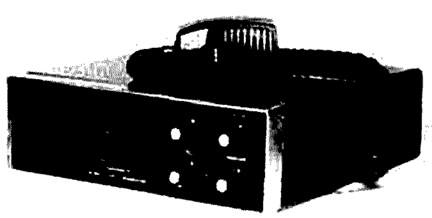
or $$P = \frac{1}{16}\,\frac{(\text{diameter})^2}{\text{depth}}$$

The units of P are the same as those used to measure the depth and diameter. This method also works for orange-peel or segment dishes, but cannot be applied directly to off-center-feed dish antennas.

John M. Franke, WA4WDL

new products

six-meter transceiver



Linear Systems has announced the introduction of a new amateur transceiver for use on the 6-meter band. The transceiver, known as the SB-50, is completely solid-state and weighs only 7 pounds. The SB-50 is synthesized with variable frequency control of both receiver and transmitter with separate receiver incremental tuning control. It covers the band from 50.05 to 50.28 MHz. The new transceiver should be especially useful for mobile installations since it contains a very effective noise limiter.

The SB-50 is rated at 20-watts PEP input and 8-watts a-m. Receiver sensitivity is less than 0.5 microvolt for 10-dB S + N/N and selectivity of 20 dB at 3 kHz and 60 dB at 6 kHz. Additional features include a lighted S-meter which indicates receive signal strength as well as power output in both the ssb and a-m mode, separately adjustable receiver incremental tuning and an external speaker connection. The transceiver comes equipped with push-to-talk dynamic microphone and mobile mounting bracket.

For further information regarding the new SBE SB-50 6-meter transceiver, write to Linear Systems, Inc., 220 Airport Boulevard, Watsonville, California 95076, or use *check-off* on page 94.

two-meter converter

Janel Labs has announced a new crystal-controlled two-meter converter that combines an impressive list of performance features with a low selling price. This new converter, the 144CC, rounds out the Janel line that already includes the deluxe 144CA high performance two-meter converter. Other products include converters for 50, 220 and 432 MHz as well as a complete line of receiving preamps.

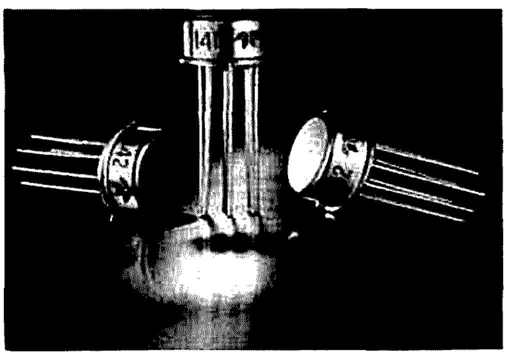
The new 144-MHz converter uses gate-protected dual-gate mosfets to provide high sensitivity while avoiding serious overload effects. The carefully designed circuit allows full utilization of the mosfet sensitivity with one rf amplifier. This is of great help in preventing cross-modulation overload by keeping the signal level low at the mixer. It allows reception of signals with 15 to 20 dB greater strength than is possible for converters wtih two rf stages.

The converter is virtually free from birdies due to the use of a seventh overtone crystal oscillator. This high overtone oscillator eliminates the need for frequency multipliers. This feature, standard in all Janel converters, is very effective in reducing suprious responses. Three tuned circuits between the rf amplifier and the mixer complete the defense against spurious responses.

More Details? CHECK–OFF Page 94

An attractive, metallic green, die-cast cabinet is used with this compact converter. BNC connectors are provided on the back panel for input and output. A power connector for 12 Vdc is also provided. Gain is 20 dB and noise figure is 3 to 5 dB. Converters are available for i-f frequencies of 26-30 MHz or 28-32 MHz. The units, completely guaranteed, are priced at an economical $49.95, postpaid. Order from Janel Laboratories, Box 112, Succasunna, New Jersey 07876. For more information, use *check-off* on page 94.

general-purpose op-amps



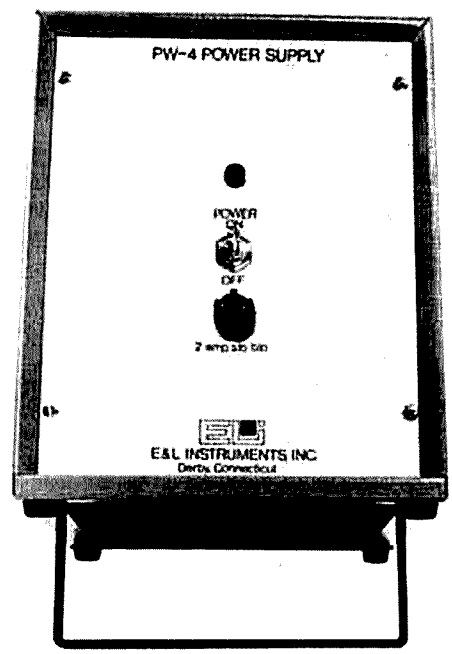
Teledyne Semiconductor has introduced a low cost, general purpose operational amplifier series, LM 141/142, which fills the performance gap between the 741 and 108 type op amps. Improved electrical characteristics of the new series include an increased slew rate of 2V/μs providing full output voltage swing through the audio frequency range and reduced input bias current of 30 mA maximum and 5 mA input offset current maximum.

The LM 141 series is fully compensated internally and is compatible with existing circuit designs using the popular 741, 107 and 1556. The uncompensated LM 142 series is a replacement for 101A, 748 and 777 applications and approaches the input performance of the 108 series amplifiers at a significant price reduction. The LM 141/142 is expected to fit applications where the 741 falls short on speed and impedance performance. They have excellent characteristics for sample

and hold circuits, long interval integraters, timers and active filter through the audio frequency range. Each amplifier meets all of the standard electrical characteristics and also includes the convenience features of short circuits and latch up protection found in other general-purpose amplifiers.

Teledyne Semiconductor will provide a free sample of the LM 141/142 to qualified engineers who respond on company letterhead with information on intended application. For more information, write to Teledyne Semiconductor, 1300 Terra Bella Avenue, Mountain View, California 94040, or use *check-off* on page 94.

base-station power supply



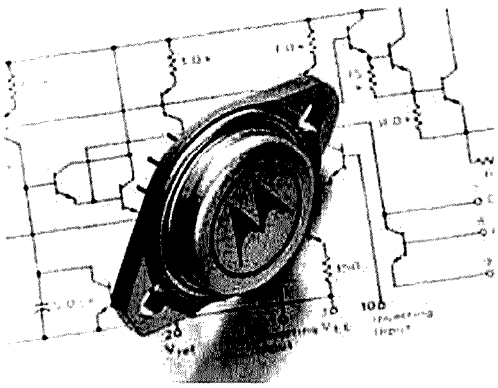
E&L Instruments has developed a new power supply designed primarily for ham radio enthusiasts. The unit, called the PW-4, produces enough power to operate both an fm transceiver and an amplifier simultaneously. The new PW-4 uses 110-120 volt ac input power, and produces a rated output of 13 volts dc at 10 amps, IC regulated to ±3%. This increased power capability means that amateurs with mobile units in their cars may take them into homes for use at night. The

More Details? CHECK–OFF Page 94

PW-4 features a modern cabinet design, current limiting and reliable heavy-duty components.

It can be used with most 12 to 13-volt dc transceivers, together with 50 to 60-watt amplifiers. The PW-4 is available direct from the factory, or from local distributors, at $84.95. For more information contact E&L Instruments, Inc., 61 First Street, Derby, Connecticut 06418, or use *check-off* on page 94.

programmable voltage regulator



A 100-watt hybrid silicon voltage regulator capable of line regulation of 0.10 percent and load regulation of 0.15 percent has been introduced by Motorola. The new MPC1000 is a 10-ampere positive or negative series voltage regulator capable of operating with input voltages as high as 60-volts. Output voltage can be adjusted from 2 to 35-volts.

Output currents of 10-amperes are easily obtained from the MPC1000 without external pass transistors; however, circuits using external pass transistors can expand the capability of the regulator to handle currents in excess of 50-amperes. Current limiting protection also has been built-in to protect the regulator from excessive surge currents.

The price for the MPC1000 in a 9-pin, metal TO-3 package is $14.95 in single unit quantities. For more information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 94.

slinky dipole antenna

The Slinky Dipole is a new Amateur antenna which combines good performance with practical size and the ability to erect or disassemble in a minimum time. It requires no matching network for low vswr operation in 50-ohm systems, and can be installed indoors in an attic or crawl space with 25- and 70-feet of available length.

The Slinky Dipole operates at 80, 40 and 20 meters and comes in kit form, including a balun, center insulator, 50-feet of RG-58/U coaxial feedline, a PL259 connector and a pair of specially made Slinky helical spring conductors. Assembly of the antenna from the kit components takes about one-half hour, and the initial setup from the completed kit to an operating antenna can be accomplished in another 30 to 60 minutes, depending on the particular installation.

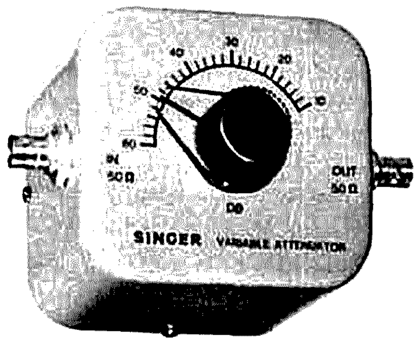
The performance of the antenna is as good as that of a full-size dipole. The power capacity of the antenna is 1000 watts CW (2000 watts PEP on ssb), and the vswr is typically less than 2.5:1 over the full 80/75-meter band and less than 1.8:1 over the full 40- and 20-meter bands.

The antenna is very versatile. As opposed to normal dipole antennas which only operate at discrete lengths, the Slinky Dipole will perform for any available length between 24' and 70' on 80/75 meters, between 12' and 35' on 40 meters, and between 18' and 6' on 20 meters. The antenna will also work in an apartment house, provided that the steel supports used in the construction are more than 150-feet apart for 75-meter operation (75-feet apart for 40-meter operation), and that the walls are essentially non-conducting.

The price for the complete antenna kit is $24.95, including all accessories, or $14.95 for the special Slinky coils alone. Include $1.00 for shipping. For more information, write to Teletron Data Corporation, 2950 Veterans Memorial Highway, Bohemia, L.I., New York 11716, or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

variable rf attenuator



Singer Instrumentation's new variable rf attenuator operates from dc to 500 MHz with an attenuator range of 10 to 60 dB. It is particularly suitable for coupling between instruments, checking transmitter out-put and receiver system degradation. Power dissipation is 100 mW. Input and output impedances are 50 ohms. Typical accuracy curves for each 10 dB of attenuation over the full frequency range are provided. The unit is priced at $130 and is available from Singer Instrumentation, 3211 South LaCienega Boulevard, Los Angeles, California 90016. For more information, use *check-off* on page 94.

eliminating engine interference

Engine interference has long been a major problem to amateur mobile operators. This new book is concerned with solving this problem in a practical manner. It explains why modern engines create interference and discusses the parts of the engine that contribute to the problem. Instructions are included on how to identify and isolate the specific components that generate noise.

Commercial noise-suppression and shielding techniques are discussed, and instructions are given for their installation. Diagrams covering the most common types of automobile ignition wiring have been included. Automatic noise limiters are also covered as are such other interference problems as instruments, wheels and tires, turn and stop signals, power-supply vibrators and antennas. 128 pages, softbound, $4.50 from HR Books, Greenville, New Hampshire 03048.

short circuits

capacitance meter

In the circuit for the capacitance meter on page 50, of the August, 1972, issue there should be a 2200-ohm resistor connected between the base of transistor Q1 and the −6 volt bus.

motorola dispatcher conversion

Inquiries and parts orders for this popular Motorola conversion should now be sent to the author at his new address: John Darjany, WB6HXU, 622 Pacific Avenue, Long Beach, California 90802.

vhf superregen receiver

In the July, 1973, issue, on page 23, the schematic for the vhf regenerative receiver should include a 0.01-μF capacitor between the wiper of the 5000-ohm gain control and the base terminal of the first TIS97 transistor.

continuous-phase audio-shift keyer

In the continuous-phase audio-shift keyer published in the October, 1973, issue the 2N5033 field-effect transistors used at Q1 and Q3 must be Fairchild's. It has been found that 2N5033s from other manufacturers have a different "on" resistance and are not usable in this application.

two-meter cavity filter

The dimensions for the two-stage, vhf cavity filter shown in **fig. 3** on page 25 of the December, 1973, issue are incorrect. Use the following corrected dimensions when building this filter.

	frequency (MHz)			
	50	144	220	432
A	41.0"	17.0"	7.30"	5.00"
B	38.0"	15.0"	5.20"	4.80"
C	35.0"	12.9"	4.00"	3.80"
D	3.0"	3.0"	3.00"	3.00"
E	4.5"	1.5"	3.00"	3.00"
F	1.4"	0.375"	1.00"	0.75"
G	3.0"	2.1"	1.20"	1.00"
H	3.0"	1.063"	2.75"	2.75"
J	0.75"	0.75"	0.75"	1.00"
K	see text			

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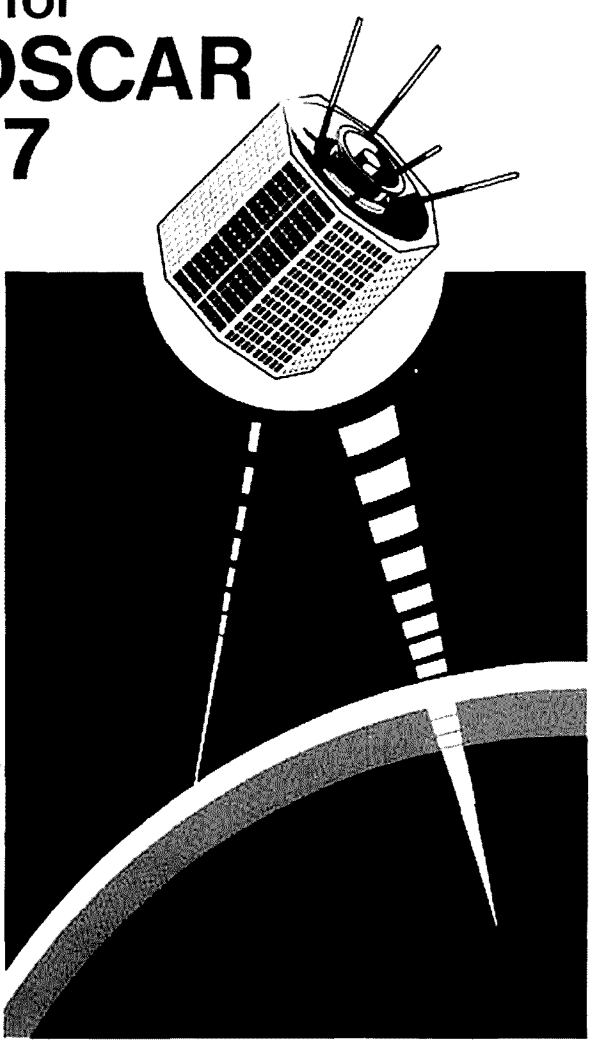

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offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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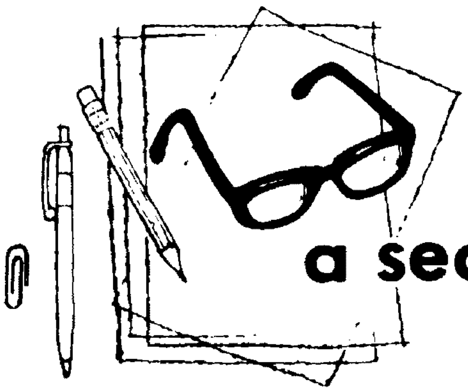


a second look

by jim fisk

It's spring and hamfest time again. No matter where you look you see an announcement for yet another convention, auction, hamfest or fm talk-in. The Dayton Hamvention in Dayton, Ohio, which is billed as the original hamvention, is one of the biggest of the year. Drawing upon a large amateur population in the Midwest, the Dayton show has provided the model for many successful amateur conventions around the country. Growing by leaps and bounds in recent years, more than 6400 hams were in attendance last year and 8000 are expected this year when the Hamvention opens its doors the last weekend in April.

This year the Dayton Hamvention Committee has gone all out to ensure a lively, interesting weekend for all. Bright and early Friday morning, the 26th of April, amateur radio manufacturers and distributors will start setting up their exhibits. At twelve noon the exhibition doors will be opened to the public. That evening the Old Old Timers and Quarter-Century Wireless Association will hold a dinner meeting in downtown Dayton.

By Saturday morning things will really start booming around Hara Arena. Vendors and traders from miles around will be setting up shop for the famous Dayton Flea Market, and the three-hour DX and vhf forums will be kicked off. A special ladies program, including luncheon, will begin at 11:30 AM, and an ARRL Forum is scheduled for the afternoon, followed by technical sessions on amateur television and troubleshooting. The traditional Saturday-night cocktail hour and banquet begins at 7:00—Senator Barry Goldwater, K7UGA, will be the guest speaker.

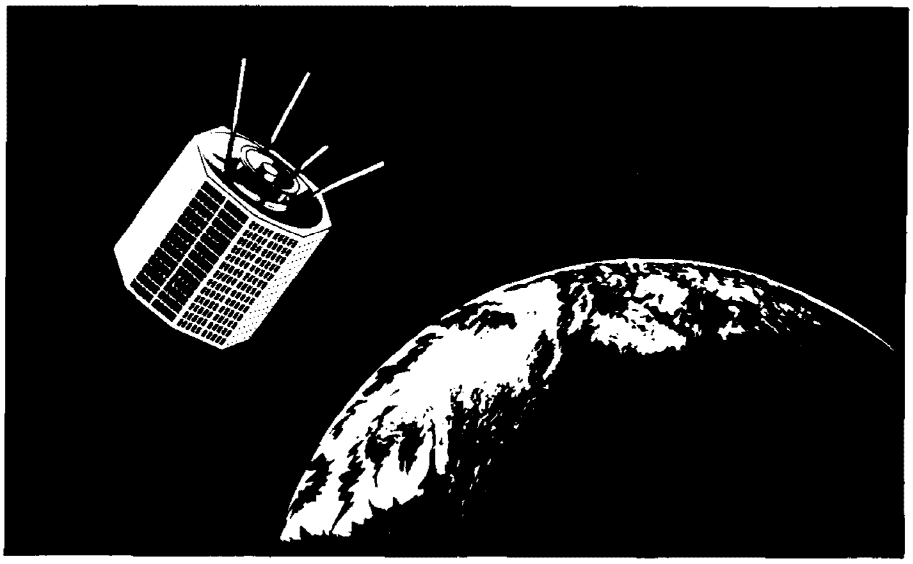
Sometime before the Hamvention is opened to the public, a 430-MHz transmitter will be hidden somewhere in the Arena area. Transmitter hunts will start at 1300 on Saturday and Sunday. If you're going to Dayton and want to join the fun, write to Rudy Plak, W8ZOF, for an antenna design.

Sunday morning the flea market and exhibit area will open at 9:00 AM, and the *antenna and state-of-the-art forums* will get under way. In the afternoon there will be forums on fm and repeaters and space communication. In addition, there will be technical and group meetings for ARPS, MIDCARS, OSSB and MARS. Other special groups attending the Hamvention are the Ohio Sideband Net, Buckeye Belles, Country Cousins, Poverty, Cracker Barrel, Firebird, Post Office, Intercontinental Traffic and Young Ladies Radio League. Prizes will be awarded at the end of each technical session on Saturday and Sunday. If past performance and the 1974 schedule are any indication, it should be another great show.

For amateurs who arrive in trailers and campers, parking will be permitted in designated areas. For those who stay at hotels downtown, free bus service will be provided out to the Arena. An allotment of 500 rooms has been set aside for the Hamvention by the local hotels and motels, so all room requests should be directed to the Accommodations Committee so that rooms can be alloted within the available supply. For more information, and a Hamvention brochure, write to the Dayton Hamvention, Post Office Box 44, Dayton, Ohio 45401.

If you've never been to the Dayton Hamvention, but have considered it, this is the year to go. If you've been before, you already know what I'm talking about. See you there!

Jim Fisk, W1DTY
editor-in-chief



communications techniques for OSCAR 7

A discussion of the new OSCAR 7 amateur radio communications satellite and the equipment for using it

Joe Kasser, G3ZCZ, 1701 East-West Highway, Apt. 205, Silver Spring, Maryland

The new OSCAR 7 communications satellite which will be launched into orbit in the near future is the most complex amateur-radio satellite built so far. It is the second in the AMSAT-OSCAR-B series of long-life amateur spacecraft, and is built in an octahedral configuration which provides surface area for enough solar cells for a positive power budget system. This means that, unlike OSCAR 6, it should not be necessary to periodically command the spacecraft into recharge modes.

OSCAR 7 will contain two repeaters and two auxiliary beacons, as well as Morse code and telemetry encoders. The two- to ten-meter repeater has an output power of two watts so signals received on the ground will be somewhat stronger than those received from OSCAR 6. The second repeater, which was built by a West German group, AMSAT-Deutschland, has an input at 432 MHz and an output at 146 MHz. The two beacons will be at 435.1 and 2304 MHz.

Ground control of the spacecraft is provided by command receivers in each repeater, redundant command decoders and a control-logic sub-system experiment. The whole spacecraft system was described in detail at the ARRL Technical Symposium on Space Communications.[1] The purpose of this article is to

present some ideas and techniques for using the new satellite for amateur communications.

working through the two- to ten-meter repeater

The ground equipment necessary for working through the OSCAR 7 two- to ten-meter repeater is identical to that required for use with OSCAR 6. Since the downlink signal will be transmitted with approximately twice the power of the OSCAR 6 transmitter, this will allow some (but not much) relaxation in the receiving equipment requirements. A ground-based transmitter with an output on the order of 80 to 100 watts effective radiated power (erp) will again be suitable. The same receiving antennas may be used.

The preferred antennas for both transmitting and receiving are simple non-directional ones such as a 5/8-wavelength vertical or a turnstile. Although the satellite's two-meter antennas are circularly polarized, it is also preferable to have some sort of ten-meter polarization diversity so you will be able to receive both vertically and horizontally polarized signals.

The passband and beacon frequencies for the OSCAR 7 two- to ten-meter repeater are slightly different from those used in OSCAR 6 (see **table 1**). These new passband frequencies were chosen for several reasons. First of all, in region 1 (Europe) 145.950 to 146.000 MHz is used by beacon transmitters operating on a 24-hour-a-day basis. These beacons are used for propagation studies, setting up converters and as a general guide to the *vhf propagation conditions* prevailing at any time. These beacons, although transmitted through OSCAR 6, provided no communications service and unnecessarily drained the power supply.

Also, in region 1 the amateur two-meter band spans only the frequencies from 144 to 146 MHz. Any transmissions above 146 MHz are non-amateur. In England, for example, the police use frequencies at 146 MHz for mobile communications and these signals were retransmitted through OSCAR 6.

In the United States, frequencies above 146 MHz are used as input frequencies for two-meter repeater installations. Stations communicating through OSCAR 6 and working DX stations on their "own" frequency above 146 MHz were also being copied through their local repeaters. The simplest solution to these problems was to move the input passband to 145.850 to 145.950 MHz.

table 1. OSCAR two- to ten-meter repeater passbands (±3 dB points).

satellite	uplink	downlink	beacon
OSCAR 6	145.900-146.000 MHz	29.45-29.55 MHz	29.45 MHz
OSCAR 7	145.850-145.950 MHz	29.40-29.50 MHz	29.50 MHz

working through the 432- to 145.9-MHz repeater

Working through the 432- to 145.9-MHz repeater will be very much like working through OSCAR 4, but in reverse. OSCAR 4 received signals on 144 MHz and re-radiated them on 431.9 MHz. The OSCAR 7 repeater receives signals on 432.1 MHz and retransmits them on 145.9 MHz. The OSCAR 7 repeater also features sideband inversion so that, for example, an upper-sideband (USB) *input* signal will be re-radiated as a lower-sideband signal. At present, many more stations are equipped to copy 144-MHz ssb than are equipped to transmit ssb on 432 MHz, and the convention is to use USB on 144 MHz. AMSAT suggests that USB be adopted as the standard for the uplink to make it possible to easily distinguish between satellite (LSB) and terrestrial (USB) signals on two meters.

The recommended transmitting power is 300 to 400 watts erp. This is best achieved by means of a high-power transmitter and a simple antenna. A ground-

plane, 5/8-wavelength whip or turnstile antenna do not require any pointing during the orbital pass, allowing the operator to concentrate on the important business of communicating.

Alternative methods of generating the required rf power are to use converted uhf fm transmitters, frequency transverters or surplus tripler-amplifiers. Since the uhf repeater is a linear device, as is the two- to ten-meter unit, the recommended modes of operation are ssb and CW.

An alternative to ssb for voice transmission is series-grid amplitude modulation, also known as controlled-carrier modulation. With this system the modulation is applied to the screen of the final amplifier tube, the carrier output is set to a low level without any modulation, and the modulation controls the level of the carrier. Thus, the louder you talk, the more power you put out. Modulation can be set to a maximum of 95%. A suitable circuit is described in reference 2 and shown in fig. 1. The S-meter will fluctuate with the modulation. This can be annoying or impressive as the case may be.

The received signal cannot be distinguished from conventional plate-modulated signals by audio means. In fact, many ssb operators will not notice that the incoming signals are a-m. You might, however, get reports of excessive carrier on your signal. I know of one G3 who was using series-grid modulation on 20 meters about four to five years ago and he received many stateside QSL cards for two-way ssb contacts!

copying the RTTY telemetry

From the telemetry point of view, the biggest difference between OSCAR 7 and the previous amateur spacecraft is the fact that OSCAR 7 has the facilities for transmitting telemetry by means of RTTY. The RTTY will be FSK on the 435.1-MHz beacon and AFSK on the 145.98- and 29.50 MHz beacons. The Doppler shift will be about ±10 kHz on the 435.1-MHz beacon, about ±3 kHz at 145.98 MHz, and ±600 Hz at 29.5 MHz. The effects of Doppler shift on the telemetry signal will be copied by the ground station as a change in the carrier frequency and not as a change in the modulation frequencies.

An extremely simple arrangement will enable good copy of the AFSK signals on

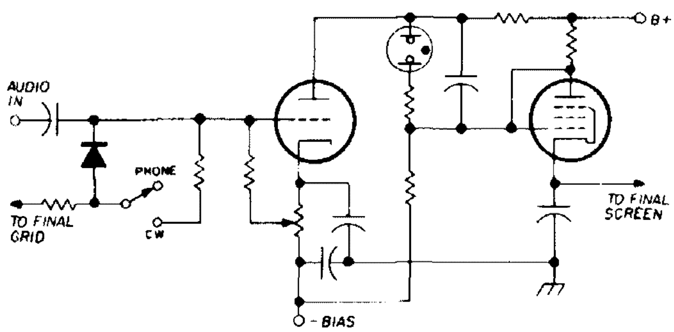


fig. 1. Basic circuit for a series-gate modulator (see text).

145.98 MHz. A typical set-up is a two-meter a-m receiver with a bandwidth of 5 to 10 kHz. A front-end preamplifier should be used between the antenna and the receiver. To copy a satellite pass it will only be necessary to tune to the low side of the signal for acquisition. The Doppler effect will cause the carrier frequency to have the appearance of a slow drift through the i-f passband of the receiver. The signal should still be within the passband at loss of signal (LOS). The detected RTTY tones do not vary during the pass so the terminal unit receives the correct tones throughout the pass. A terminal unit such as the ST-5 (see reference 3) would be very suitable for this application.

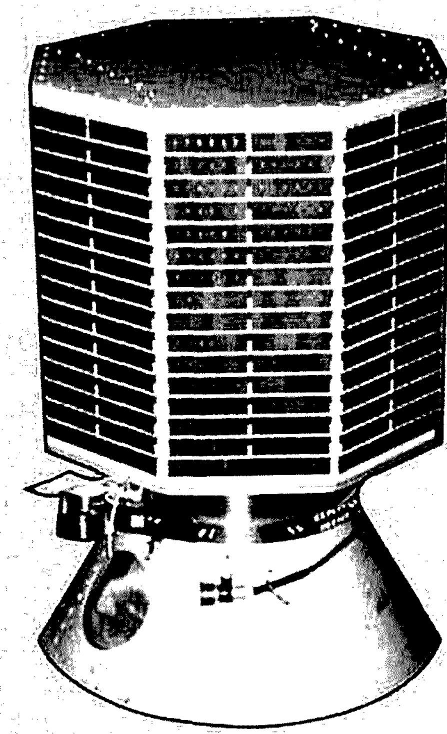
Copying the FSK transmissions on 435.1 MHz will be slightly more difficult. Tests were conducted from club station WA3EWJ in March, 1973, to determine the feasibility of copying satellite FSK RTTY signals with simple ground equipment, and to investigate how Doppler shift would affect the received signals. It was found that copying the teletype signals was quite easy. The terminal unit used was the ST-5.

The Doppler shift on the signal was found to appear as a gradual drift downward in frequency throughout the pass.

The received signal was monitored using the conventional cross-shaped oscilloscope pattern, and the vfo on the receiver was adjusted to keep the display correct. Frequency-selective fading was also observed on the signals. Although the ST-5 provided good copy (even after two tape transfers), it was necessary to stay at the receiver to adjust the vfo every minute or so. It would be much nicer if the receiving system could tune itself, automatically tracking the Doppler shift, freeing the operator for other tasks.

Two designs for automatic frequency control of received RTTY signals have been published in recent years. Reference 4 describes a terminal unit using a phase-locked loop. The use of such a terminal unit allows the signal to drift through the passband while printing out good copy. However, when the signal moves out of the passband, copy is lost.

Reference 5 describes an add-on unit for the ST-5 or ST-6 terminal units which consists of a circuit which monitors the mark frequency and puts out a proportional positive or negative voltage (with respect to ground) as a function of how far (and which way) the receiver is tuned off the signal. If a center-zero meter is connected across this afc output, it will act as a tuning meter, showing how the receiver is tuned with respect to the incoming RTTY mark signal.



The AMSAT-OSCAR-B communications satellite which will be orbited as OSCAR 7 later this year. In this photograph the spacecraft is mounted on a sheet-metal cone which is installed in the launch vehicle. The OSCAR package is separated from this cone and ejected into a sun-synchronous orbit approximately 900 miles above the surface of the earth. The antennas, which are stainless steel tapes, are unfurled after the spacecraft goes into orbit. OSCAR 7 will carry 145- to 29-MHz and 432- to 146-MHz repeaters as well as beacons on 435.1 and 2304 MHz.

Reference 5 also describes how to modify the receiver vfo to accept and use the automatic frequency control signal. An alternative approach is to build a new external vfo for use with OSCAR 7. If a new vfo is to be built, a better approach is to build an afc-controlled oscillator for the front end of the 435.1-MHz converter.

copying the s-band beacon

Link calculations for a typical receiving station for the 2304-MHz beacon were given in a previous article.[1] It was shown that with a Doppler shift of about ±55 kHz, a receiver bandwidth of 500 Hz and a four-foot dish with a pointing

accuracy of ±7.5°, reception of this beacon presents a real challenge.

A simple front end converter for 2304 MHz is described in reference 6. It uses the classic trough-line front end based on earlier 1296-MHz units. Antenna construction plans are given in reference 7 in an article describing a pulse communication system. Amateurs who are currently tracking OSCAR 6 in both azimuth and elevation using a narrow-beamwidth antenna should already have the ability to track OSCAR 7 with a four-foot dish. Thus, reception of this beacon is not quite as difficult as it appears at first glance.

using medium-scan television

In most countries, wideband TV is an authorized mode of transmission in the 432-MHz amateur band. Since OSCAR 7 contains a repeater having an input in the 70 cm band it opens up the prospects of live, long-distance, real-time TV communications.

However, since the signals are reradiated on 145.9 MHz, a waiver or special permit must be issued by the licensing authorities. This permission has already been requested of the FCC. Also, since the repeater has only a 50-kHz passband the transmissions will still be limited in bandwidth. This rules out standard fast-scan (525/625 line) pictures. Slow-scan TV with its eight-second frame rate is suitable (as demonstrated by a number of OSCAR 6 contacts) but could be improved upon, at least with respect to the frame rate.

Reference 8 describes a medium (or faster slow-scan) TV system. The specifications for this system are such that all frequencies used are four times the equivalent normal slow-scan rate as shown in **table 2.** The pictures thus have a frame rate of 2 seconds. These TV pictures are not currently used for on-the-air transmissions because the bandwidth is also four times the normal slow-scan TV bandwidth. The pictures are used for setting up cameras and monitors in the home station and are converted to slow-scan TV by the very simple method of dividing all frequencies by a factor of four (using digital ICs). The pictures are then indistinguishable from conventionally generated sstv.

table 2. Operating parameters of slow-scan and medium-scan television signals.

parameter	sstv	mstv
Line rate	15 Hz	60 Hz
Number of lines	120	120
Frame time	8.1 sec	2.025 sec
Sync frequency	1200 Hz	4800 Hz
Black frequency	1500 Hz	6000 Hz
White frequency	2300 Hz	9200 Hz
Horizontal sync pulse	5 ms	1.25 ms
Vertical sync pulse	30 ms	7.50 ms
Video bandwidth	900 Hz	3600 Hz

Since the bandwidth of medium-scan TV (mstv) is less than 15 kHz, and the spacecraft repeater has a 50-kHz passband, there does not appear to be any reason why mstv should not be used as a communications mode through OSCAR 7, providing that the relevant permits are issued by the licensing authorities. With its two-second frame time, mstv is a vast improvement over sstv eight-second frame time.

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ham radio

simple active filters for direct-conversion receivers

How to design simple active audio filters for radio communications

Wes Hayward, W7ZOI, 7700 SW Danielle, Beaverton, Oregon 97005

Many of the simpler communications receivers in use today are lacking in selectivity for CW work. To some extent this deficiency can be minimized by using audio filters. The more interesting designs, at least for me, are those where active RC circuits are used to replace the classic LC configurations. While high quality active filters are available commercially,* they are also popular as projects for the amateur experimenter. Such filters are especially useful in conjunction with direct-conversion receivers, an approach in which all adjacent channel selectivity must be obtained at audio frequencies.

Several of the active filter designs available to the experimenter suffer from problems which make them less than optimum for general use. For example, many of the designs are aimed at achieving rather narrow bandwidths, often less than 100 Hz. While these units are quite useful for some specialized applications, I prefer a somewhat wider bandwidth for general CW work. A 0.5-kHz wide response is usually more than adequate if steep skirts are maintained.

Many of the popular active filter designs require tight control of *component* tolerances, leading to difficulty and excessive expense when being duplicated. The work described in this article is aimed at designs which use ±10% or even 20% tolerance components and feature wider bandwidths while maintaining sufficiently steep skirt response to be useful.

*Such as those manufactured by MFJ Enterprises, Post Office Box 494, State College, Mississippi 39762 (see *ham radio,* November, 1973, page 68).

lowpass filter

Shown in fig. 1 is an abbreviated schematic of a 10-pole peaked lowpass filter I built. Although only two lowpass sections are shown, the filter contains five identical sections. Each of these has a 2-pole lowpass response which is peaked at the cutoff frequency. A similar response is that of the pi-network in a tube transmitter, again a peaked lowpass filter.

The Q of each active filter section is about 1.9 which yields a net 6-dB bandwidth of about 200 Hz. Skirt response, however, is not lacking. With a center frequency of 540 Hz, the attenuation is 75 dB at 1200 Hz. Net gain of the system is 28 dB at resonance. A single-pole highpass section is used at the input to bias the following stages and to provide some additional attenuation at the low frequencies.

The measured responses for one, three and five lowpass sections are plotted in fig. 2. This filter was built with 10% capacitors and resistors. However, the low Q of each pole pair would allow the use of 20% components with a minimal degradation in performance. Indeed, the slight stagger-tuning effect that could result might be quite desireable. The npn-pnp feedback transistor pairs are used as unity gain amplifiers and are not critical as to transistor type. The bias of the input section was chosen to compensate for the 0.6-volt offset introduced by each feedback pair, placing the filter output at half the power supply level.

From a practical point of view, the filter has been found to be an excellent performer. The moderately wide bandwidth makes the unit easy to use, even

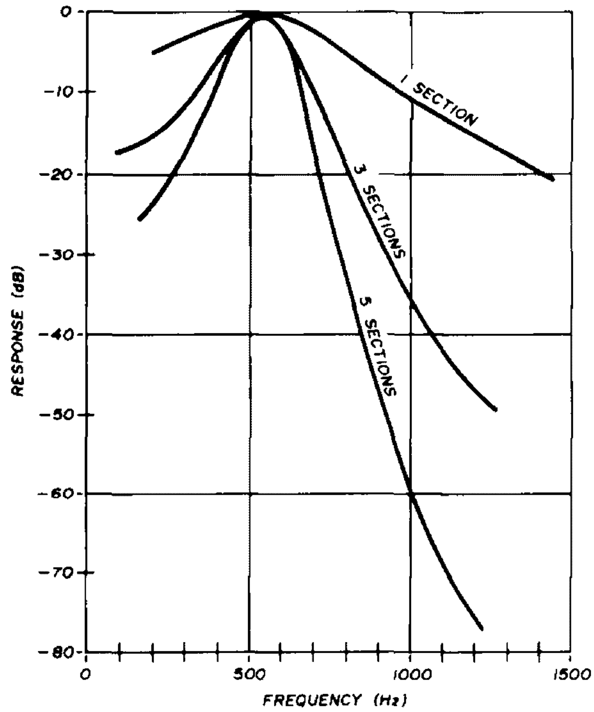
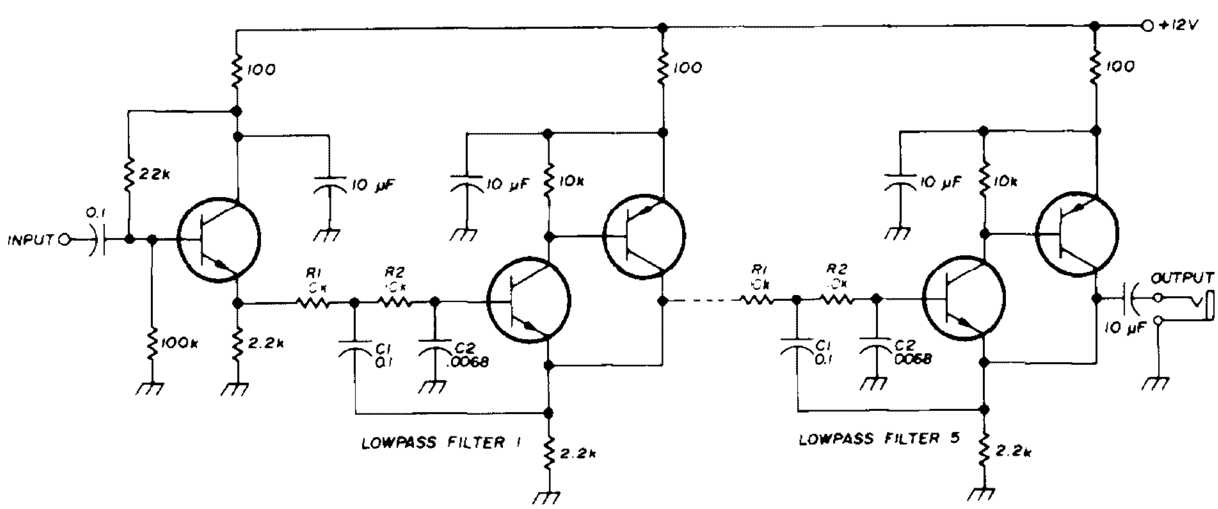


fig. 2. Measured response of one, three and five lowpass sections of the 10-pole active filter shown in fig. 1.



C1	0.1 µF, 10% tolerance
C2	0.0068 µF, 10% tolerance
R1,R2	10k, 1/4-watt, 10% tolerance

fig. 1. Abbreviated schematic of a 10-pole peaked lowpass filter. All five lowpass sections are identical. Npn transistors are 2N3565, 2N3904 or similar; pnp transistors are 2N3638, 2N3906, etc.

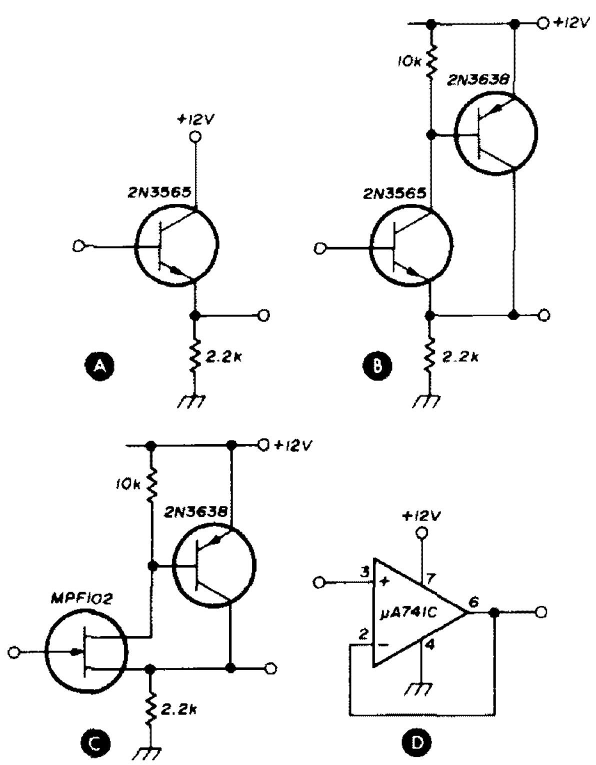
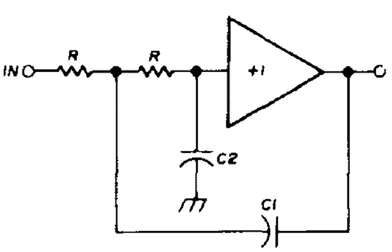


fig. 3. Four different types of unity gain, non-inverting amplifiers suitable for use in active filters.

with receivers with minimal bandspread. However, the steep skirts insure adequate rejection of adjacent channels. When used with even the most simple direct-conversion receiver, performance is suitable for the majority of amateur communications.

other filter designs

The design outlined above should be suitable for those wanting a circuit to duplicate. However, the low cost of modern semiconductors and the ease of construction of audio circuits make active filters a very attractive area for further experimentation. The remainder of this article will present some possible variations for you to try in your own lab. If you have an analytical bent you will find the analysis of the peaked lowpass design to be straight forward by using either classic methods[1] or the real-time approach outlined earlier.[2]



$$f_{CO} = \frac{1}{2\pi R\sqrt{C_1C_2}}$$

$$Q = \frac{1}{2}\sqrt{C1/C2}$$

fig. 4. Basic peaked lowpass filter section. The cutoff frequency, f_{CO}, and Q are defined by the values of R and C as shown here.

For the simple lowpass filter designs, the amplifier should have a unity, non-inverting gain and should exhibit good impedance-transforming properties. In many cases, a simple emitter follower using a high beta transistor will suffice. Integrated-circuit operational amplifiers such as the popular µA741 are excellent, although a lower noise type such as the LM-301 is sometimes preferred. Shown in **fig. 3** are four possible configurations, all suitable for use with a single power supply. The fet input amplifier is useful for low noise applications.

The basic peaked lowpass filter design is summarized in **fig. 4.** Note that the Q of the circuit is completely defined by the ratio of the two capacitors. When designing a filter, a Q is chosen and convenient capacitors of standard value are then picked. Miniaturization and low cost would suggest choosing relatively low capacitor values. However, noise considerations often point toward the use of somewhat larger values. Once the capacitors are chosen, the frequency of the unit is determined by calculating the proper resistance values. Clearly, a tunable filter with a constant Q would result form the use of a dual potentiometer.

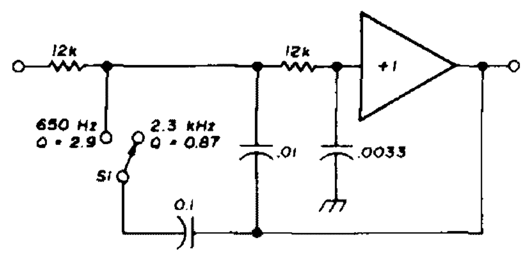


fig. 5. Dual bandwidth active filter. With switch S1 open, cutoff is 2.3 kHz; cutoff is 650 Hz with switch closed.

The filter shown will have unity gain at dc and a voltage gain equal to the Q at the "cutoff" frequency. The measured peak frequency will be slightly lower than the "cutoff" frequency defined by the equation of **fig. 4.** This error is largest at

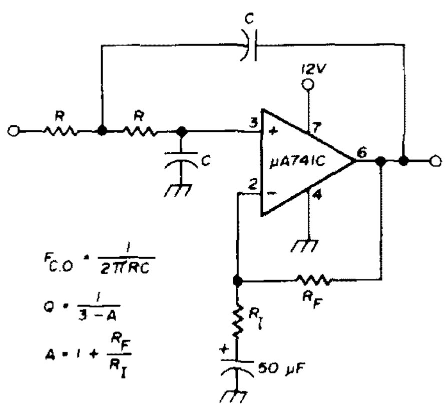
low-Q values. For example, the measured center frequency of the 10-pole filter of **fig. 1** is 10% lower than that calculated. This effect is characteristic of any low-Q resonator.

The fact that both frequency and Q are dependent upon the capacitors can be used to advantage. Shown in **fig. 5** is a dual bandwidth filter. With the switch open, the cutoff is 2.3 kHz and Q is 0.87. When the switch is closed, the center frequency drops to about 650 Hz and Q increases to 2.9. A multiplicity of these sections should yield a filter useful for both CW and ssb work.

Obtaining electronic components is often a problem for amateurs, with precision capacitors being especially difficult to find. Occasionally, you will come across a large number of capacitors of identical value. These can be used in filter applications in conjunction with non-inverting amplifiers with a gain greater than unity. This lowpass filter configuration is presented in the schematic and equations of **fig. 6**. You can see that Q can become infinite for a closed loop gain, A, of only 3. This oscillating condition could be used to advantage in a transceiver application by using an fet or bipolar switch to alter stage gain, providing a simple sidetone function.

limiting amplifier

One drawback of many simple direct-conversion receivers is the lack of agc.



$$F_{C.O} = \frac{1}{2\pi RC}$$

$$Q = \frac{1}{3-A}$$

$$A = 1 + \frac{R_F}{R_I}$$

fig. 6. Alternate form of peaked lowpass filter using a non-inverting amplifier with gain greater than one. Cutoff frequency, f_{co}, and gain, A, are defined by the resistance and capacitance components; Q is determined by gain.

This deficiency could be minimized for CW reception by careful application of limiting. For example, shown in **fig. 7** is a simple inverting, limiting amplifier with an adjustable limiting threshold. Below the threshold, the amplifier is linear with a voltage gain of 10. However, as soon as

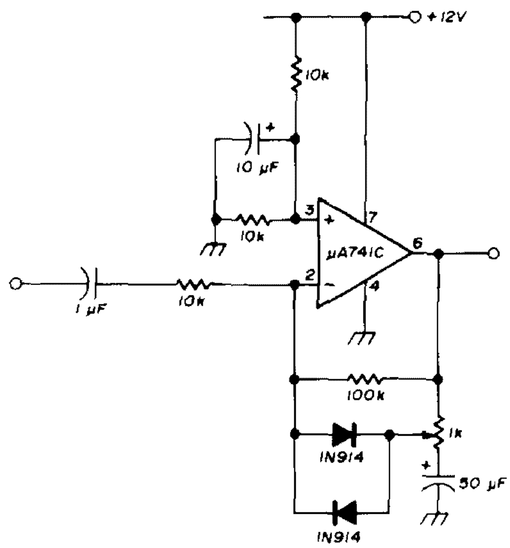


fig. 7. Simple inverting, limiting amplifier with adjustable limiting threshold.

the output is high enough for the silicon diodes to conduct, the gain drops below unity. This amplifier should be preceded by several sections of filtering, and followed by a single-section lowpass filter to eliminate the harmonic distortions generated in the limiting process. The resulting clipper-filter system would be a real ear saver when used with any receiver with poor or non-existent agc.

summary

There are many possible circuit configurations for active filtering and an equally wide variety of applications. Here we have considered only a very simple family of circuits. Perhaps this article will stimulate others to try their hand at this intriguing area.

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ham radio

telefax transceiver conversion

Modifying the Model 6500A fax machine for F5 emission

Fred J. Steurer, KØOMR, 11025 Patsy Drive, St. Louis, Missouri 63123

Hams have long been an imaginative and ingenious bunch, taking older commercial equipment and converting it for their own use. Lately the Telefax Model 6500 facsimile transceivers have been on the market for 10 - 15 dollars each. Articles have been written on the conversion of these machines, but none have been very satisfactory as many of you may have discovered.

The machines were originally designed to work over landlines in conjunction with another setup located in a central office. This setup provided the sync detection and drum feed control for a pair of machines. The machine is basically an A4 emission device; that is, a varying amplitude of the same frequency. One of the difficulties of A4 is the constant need for riding gain on the volume control to overcome band fading. Another problem exists in the legality of feeding this A4 emission into a 2-meter fm transmitter, which then makes F4 emission. This is not permitted under current regulations.

Included here are a set of electronics and standards to convert the final emission to F5, which is legal on 2-meter fm and which solves all the problems of fading on the low bands. The standards are simple; 2500 Hz for white, 2000 Hz for black, and the gray scale falls in between. The sync is a series of beeps at whatever frequency is coming out of the machine at that time. The first portion of the scan may be black, white, or gray – it makes no difference for the sync. Also included is an automatic drum feed. The receiving circuit detects the beginning of a picture and causes the drum to automatically begin the horizontal travel.

The schematic of **fig. 1** shows the unmodified Model 6500 fax machine. The following discussion describes the circuitry for adapting the machine for F5 emission and includes the sync and drum feed control.

interface circuits

Q5 and Q6 make up the multivibrator, which oscillates between 2500 Hz, the white frequency, and 2000 Hz, the black frequency (**fig. 2**). Q7 is the modulator, which detects how much output is com-

ing from the fax machine for a given picture and changes the oscillator frequency. Q8 is a simple emitter follower isolation. U2 is a flip-flop with a canceling input (**fig. 3**). On receive the phasing contact triggers this flip-flop, which drives Q3. Q3 in turn, drives PH-1, an LED lamp shining on a photo-resistor. This circuit provides isolation for the triac circuit, SCR 1. (Construction of PH-1 is shown in **fig. 4**.) Q4 is the triac control, which turns off the gate on the triac (SCR 1) when the phasing contact is triggered by the rotating drum. The triac interrupts power to the gray-colored motor (see **fig. 1**) to slow the rate of the revolving drum. When the received signal (beeps) coincides with the local phasing contact opening, the flip-flop is canceled and the gray-colored motor runs at normal speed. Both units (transmitting and receiving machines) are in sync. That is, the red line on each machine is in the same angular position.

U1 (**fig. 5**) is a standard limiter, which removes all a-m from the signal and causes a constant level regardless of the volume setting. The input is set for 500 ohms; and it would be a good idea to install a small transformer, 3.2 to 500 ohms, between the speaker terminals of the receiver and the 500-ohm input to match impedances.

The video detector is just a tuned trap in series with the limiter output and the fax input. Q1 and Q2 are the automatic drum feed and tuning circuit (**fig. 6**). Three seconds after video is detected, the relay closes and the drum feeds on the receiving end. U3 is a series-type regulator in the plus side of the power supply, **fig. 7**.

telefax transceiver conversion

Remove the top cover and bottom plate and check the tubes. Check the stylus and replace if necessary with a piece of carbon-steel wire from a wire brush.

Carefully remove the exciter lamp, lens telescope, and projection tube. Remove the lenses from these assemblies and clean them. Replace the lenses in the same order and in the same direction as removed. Remove the phototube, clean, and replace. Plug in the 117-volt line cord, and push the outgoing pushbutton. This turns on the lamp. Focus the light spot on the drum by moving the telescope back and forth. Put a piece of paper with typed letters on the drum, and focus the image on the pinhole of the projection tube by moving the projection tube back and forth. Take good care on this step if you are to transmit sharp pictures. Check that the red line on the drum is at the stylus position when the phasing contact is open and the free slack is taken up in the normally rotating direction. The factory setup may have slipped. Adjust with an Allen wrench. Burnish the phasing contacts and adjust them for 0.020 inch with a feeler gauge.

1. Clip the 51-ohm resistor (**fig. 1**) from the incoming switch on the front panel and the two other wires from this switch. This frees a set of contacts for future use. See **fig. 8**.

2. Clip the wire coming from relay LR, the normally closed contact, and going to relay HR, the moving contact.

3. Clip the wire on the rear, outer terminal of the out-going switch and run a wire from this contact to the moving contact of relay HR just made available. See **fig. 8**.

4. Clip the wires on all three lugs of the BR relay. Fold back and disregard.

5. Clip the two green wires from one side of the coil of the BR relay, fold back and disregard.

6. Connect a wire from the N.C. contact on LR, made available earlier, to the coil terminal just made available on the BR relay.

7. On the line transformer located under the gray-colored motor, cut all three wires from the terminals located on the transformer.

8. With a wire, ground the terminal closest to the chassis. A terminal lug is close by to solder to ground. See **Fig. 8**.

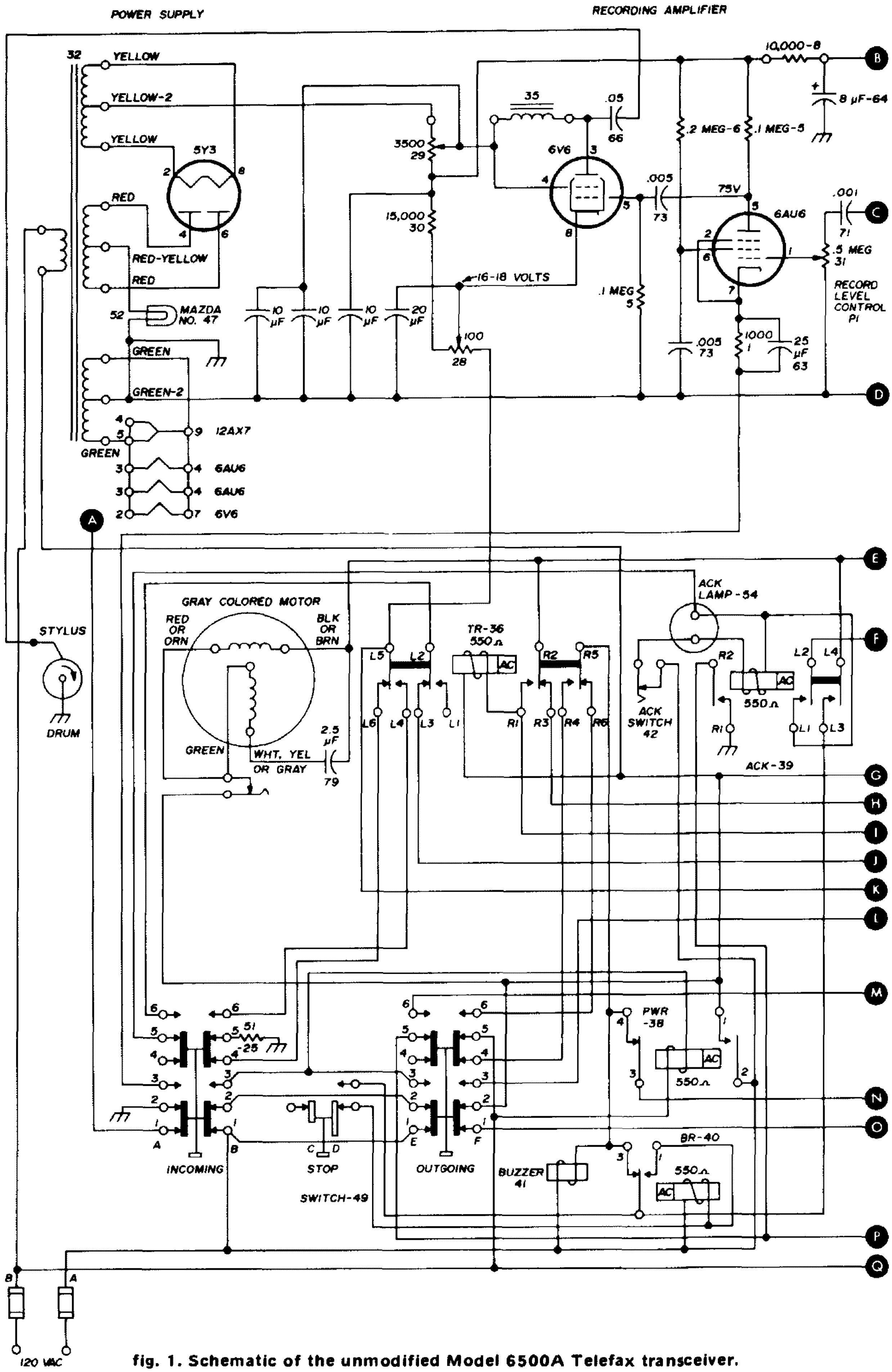
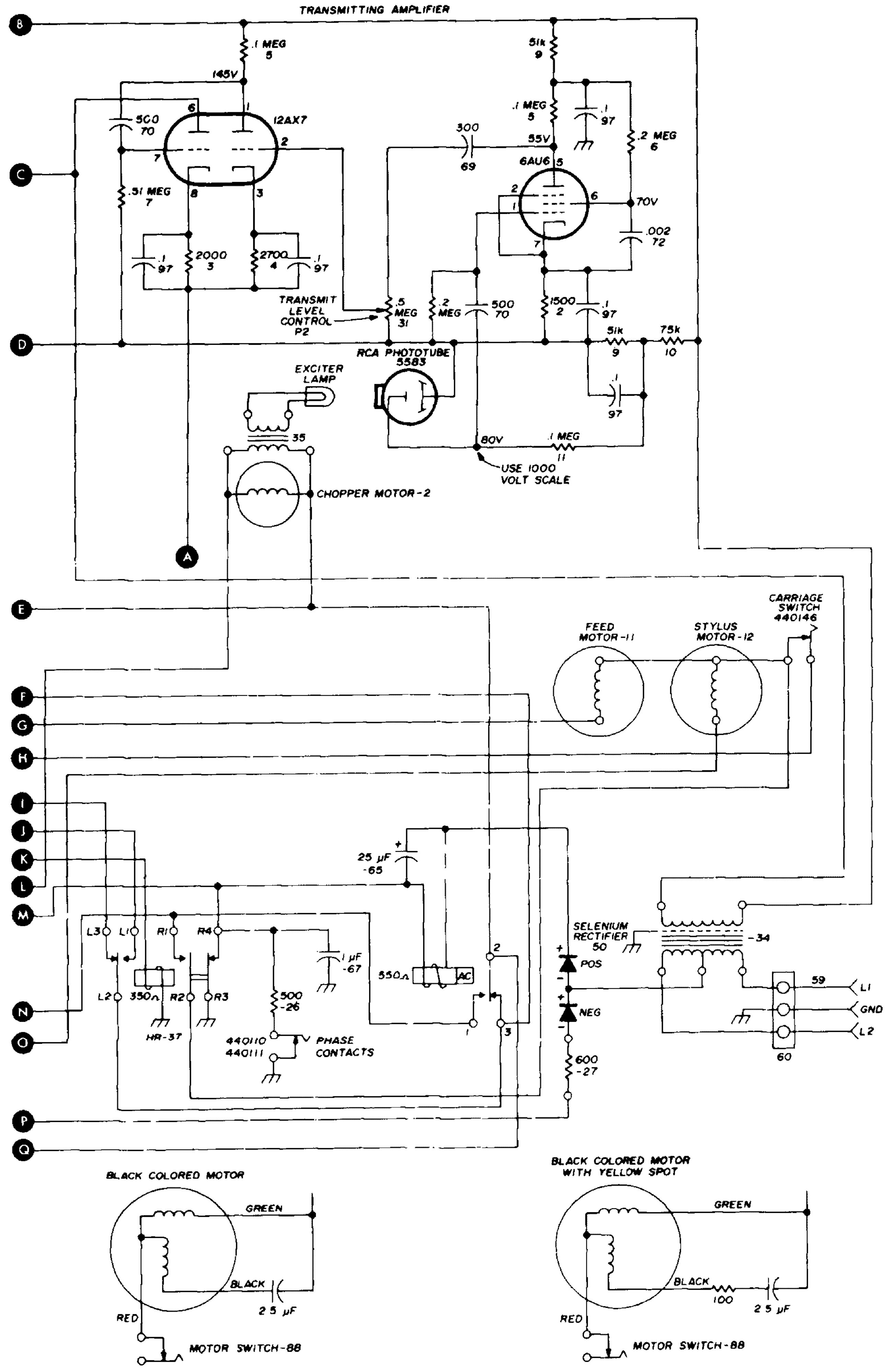


fig. 1. Schematic of the unmodified Model 6500A Telefax transceiver.


TRANSMITTING AMPLIFIER
.1 MEG 5
145V
12AX7
500 70
.51 MEG 7
2000 3
2700 4
51k 9
.1 MEG 5
.2 MEG 6
300 69
55V
6AU6
70V
.002 72
1500 2
TRANSMIT LEVEL CONTROL P2
.5 MEG 31
.2 MEG
500 70
RCA PHOTOTUBE 5583
51k 9
75k 10
EXCITER LAMP
35
80V
.1 MEG 11
USE 1000 VOLT SCALE
CHOPPER MOTOR-2
FEED MOTOR-11
STYLUS MOTOR-12
CARRIAGE SWITCH 440146
25 μF -65
1 μF -67
550Ω
AC
350Ω
HR-37
500 -26
440110 440111
PHASE CONTACTS
SELENIUM RECTIFIER 50
POS
NEG
600 -27
-34
59
L1
GND
L2
60
BLACK COLORED MOTOR
GREEN
BLACK
2.5 μF
RED
MOTOR SWITCH-88
BLACK COLORED MOTOR WITH YELLOW SPOT
GREEN
BLACK
100
2.5 μF
RED
MOTOR SWITCH-88

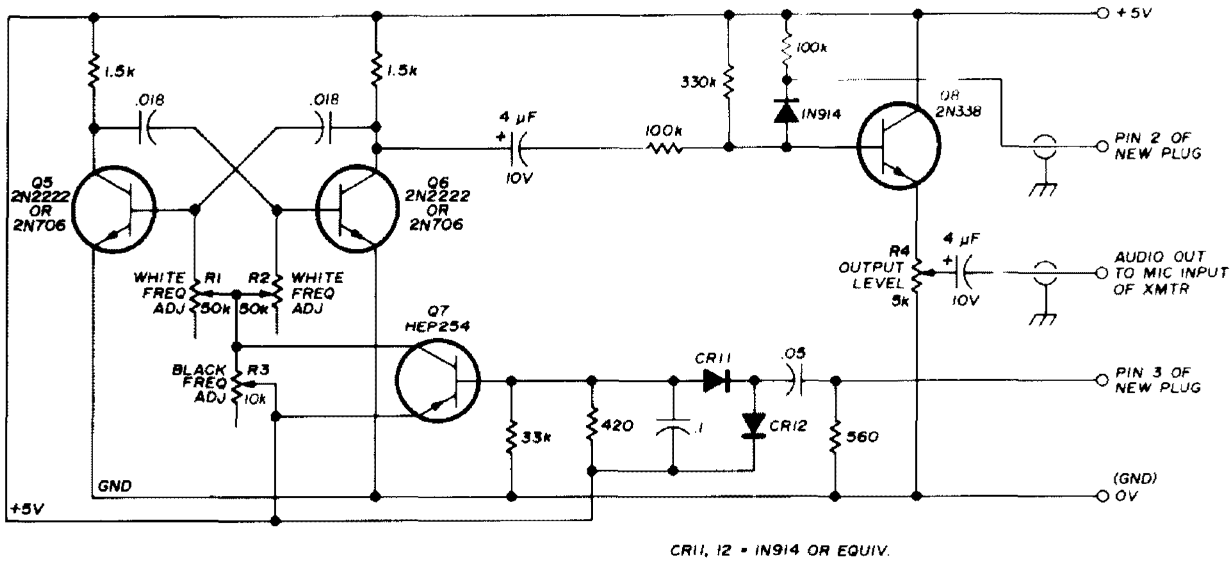


fig. 2. Modulated oscillator.

9. Clip the two wires (one from each lug) from the rear, inner spdt switch of the outgoing pushbutton and fold back the wires. See **fig. 8.**

10. To the center lug of these three, connect a wire to the last lug made available on the line transformer. No connection is made to the center tap on the line transformer. See **fig. 8.**

11. Extend the red wire (that was clipped from the line transformer) to the rearmost of the three contacts.

12. Extend the white wire (that was clipped from the line transformer) to the frontmost lug of the three. This transfers the line transformer from receive to transmit with the pushbutton.

13. Clip the two gray wires from the other coil terminal of the BR relay. Solder together and tape.

14. Connect a wire from this coil terminal just made available on relay BR to the N.O. contact, bottom stack of the PWR relay. There are several wires connected to this terminal of the PWR relay.

15. Make available a set of contacts on the TR relay as follows: On the moving contact of one of the sets is a pair of wires, one of which goes to the coil of the HR relay, the other to the 100-ohm, 10-watt resistor bolted to the chassis. After identifying this set of contacts, clip the pair off the moving contact of the TR relay, solder them together, and tape.

16. Clip the other wires from this set of contacts, fold back and disregard.

17. Clip the wire on the drum phasing contact and remove.

18. Connect a piece of shielded wire from the phasing contact to the moving contact on relay TR just made available.

19. Connect another shielded wire from the N.O. contact of relay TR to the moving contact of relay BR made available earlier.

20. Clip the blue wire on the coil of the ACK relay going to the neon lamp, fold back and disregard.

21. Connect a wire from this terminal on the coil of the ACK relay to the N.O. contact of the PWR relay. (There are several wires on this terminal.)

22. Clip both wires from the acknowledge pushbutton on the front panel.

23. Remove the switch and replace with a spst, normally open pushbutton or similar momentary-contact switch. Wiring of this switch is described later.

24. Clip three wires on one side of the neon lamp holder. One side of the neon has a wire going to the acknowledge pushbutton. If this is the side you clipped first, identify this wire and discard it.

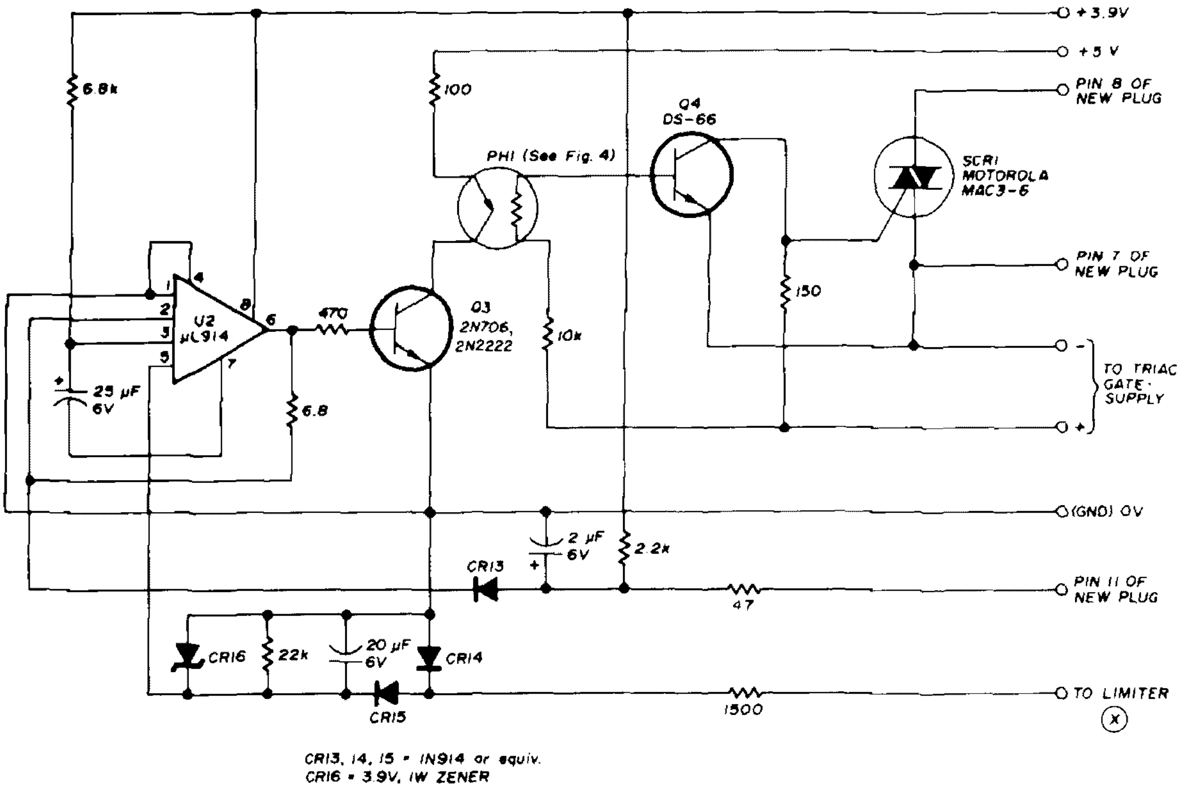


fig. 3. Sync circuit.

25. Solder the other two wires to one side of the new switch.

26. Clip three wires from the other side of the neon lampholder, solder together and tape.

27. Connect the neon lamp across the N.O. contacts on relay LR.

28. Connect the second terminal of the new switch to the moving contact of the LR relay. This wire can be picked up on the neon lamp.

29. Clip the wires from the bottom two sets of N.O. contacts and moving contacts of the ACK relay, fold back and disregard.

30. Clip both wires from the coil of the LR relay, fold back and disregard.

31. Connect a wire from the moving contact of relay LR to one terminal of the coil of relay LR.

32. Connect a wire from the other coil terminal of relay LR to moving contact of set "A" just made available on the ACK relay.

33. Connect a wire from the N.O. contact of set "A" on the ACK relay to the new wire installed earlier to the coil of the ACK relay.

34. Connect a wire from the terminal having the gray wire (from the new switch) to the moving contact of contact set "B" on the ACK relay.

35. Connect a wire from the moving contact of the LR relay to the N.O. contact of contact set "B" on the ACK relay.

36. Enlarge the hole in the rear apron to accept an 11-pin receptacle. Wiring of the new socket is as follows.

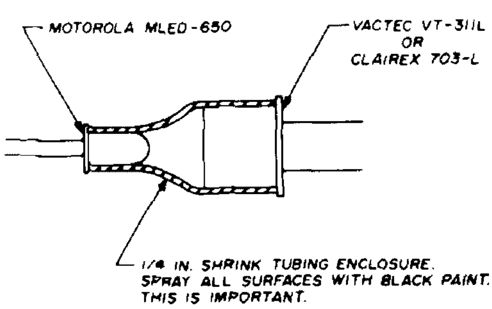


fig. 4. Construction of the PH-1 LED/photoresistor assembly. If the light-emitting diode does not operate, reverse the connections to it.

37. Pin 1 - Chassis ground.

Pin 2 - Shielded lead to N.O. contact of relay BR.

Pin 3 - Connect a wire to the *red wire on terminal* block for the line transformer.

Pin 4 - Connect a wire to the *white* wire on terminal block for the line transformer.

Apply a source of 2500-Hz voltage to the 500-ohm input to the limiter. Connect a scope or a VU meter to pin 4 of the new plug (see **fig. 5**). Put a receiving blank on the drum. Push the INCOMING switch. Adjust L1 for a minimum reading on the scope or VU meter. Push the STOP switch. Put a new receiving blank on the drum. Feed 2000 Hz into the 500-ohm input, and push the INCOMING switch. When the neon light lights, push

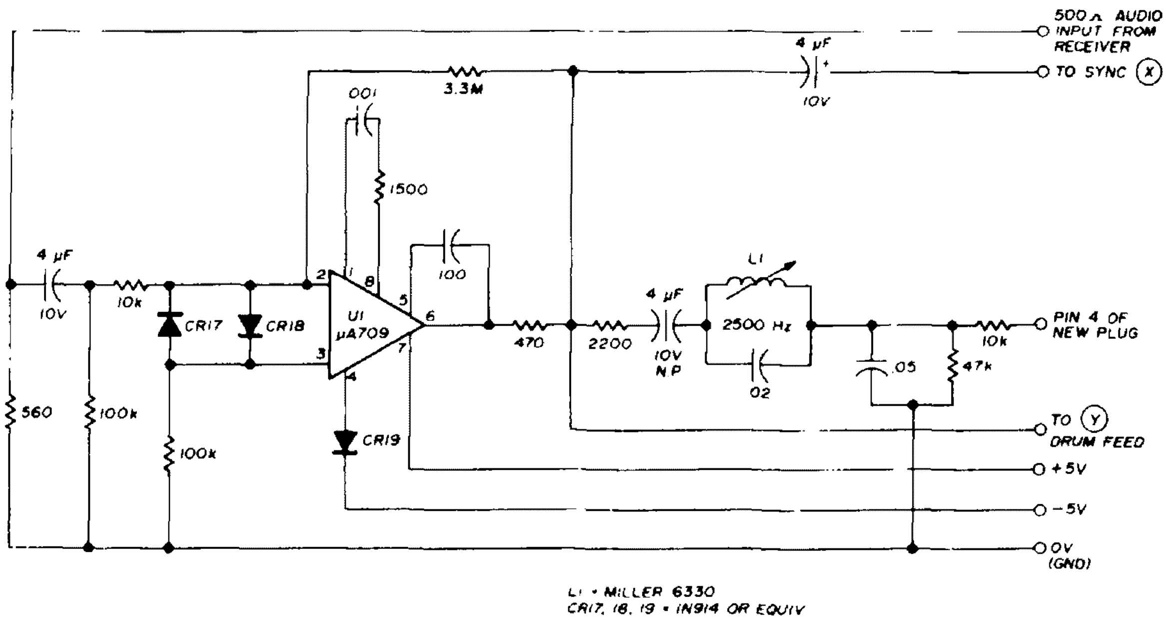


fig. 5. Limiter and video detector.

Pin 5 - Connect to the N.O. contact, "A" set, of the ACK relay.

Pin 6 - Connect to moving contacts, "A" set, of ACK relay.

Pin 7 - Clip wire to service switch (chassis) and extend to Pin 7.

Pin 8 - Connect a wire to the service switch terminal just made available.

Pins 9 and 10 - Connect one pin to each side of power transformer primary.

Pin 11 - Connect a wire to the N.O. contact of the set made available on the TR relay.

tune-up

Push the OUTGOING button and check all the voltages on the board: +5V, -5V, and +3.9V. Push the STOP button.

the drum feed pushbutton (the new switch added in the machine conversion).

Adjust P1 on the fax chassis for a good black burn on the paper. Change the oscillator to 2500 Hz. Touch up P1 just so the burn disappears. As you go from 2500 down to 2000 Hz in steps, you will see the gray scale. No further adjustment is required on the video detector. Push the STOP button.

Connect a scope to the audio output of the modulated oscillator (see **fig. 2**). Connect a frequency counter or some means of determining 2500 and 2000 Hz reasonably accurately. An audio oscillator may be used as a BFO while listening with an earphone connected at the same point. Be as accurate as possible.

Put a sending blank on the drum. With a screwdriver, turn off the service switch

located near the gray-colored motor on the top side of the chassis. Rotate the drum manually until the phasing contacts are open (next to the drum gear). Note the drum is free wheeling on the shaft for almost a full turn.

Push the OUTGOING button. The drum should not be rotating because the service switch is turned off. While the scope). After several trials the proper reading will exist.

Now, with P2 fully counter clockwise, rotate the drum to the whitest portion of the sending blank. Advance P2 until the counter just levels off and no further. The counter should read 2500 Hz. Rotate the drum to the black letter again and the counter should read 2000 Hz. Anything

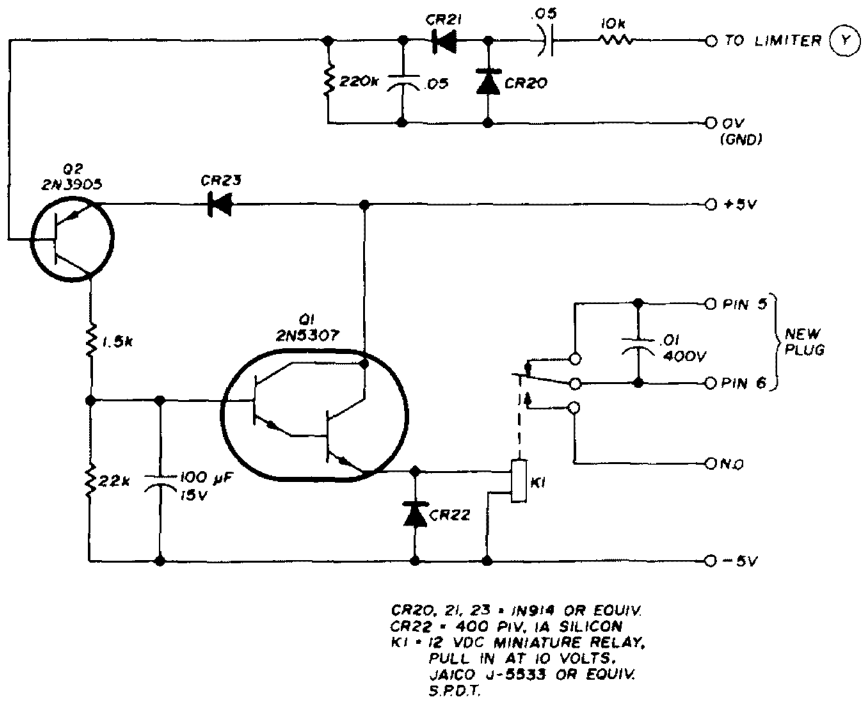


fig. 6. Automatic drum feed.

phasing contacts are still open, rotate the drum to put the light beam on the whitest part of the sending blank. Adjust the SENDING pot P2 on the machine chassis for approximately ½ turn. Adjust the two white frequency adjust pots, R1 and R2 (see **fig. 2**) for 2500 Hz on the counter. Juggle the two pots (R1 and R2) for a symmetrical trace on the scope. Now rotate the drum to a black letter on the sending blank, keeping the phasing contacts open. Turn P2 on the chassis fully counter clockwise. Adjust R3 (see **fig. 2**), the black frequency adjust, for a reading of 2000 Hz on the counter. Readjust R1 and R2 again because these adjustments interact somewhat (P2 to ½ turn, drum on the whitest portion of the sending blank, a reading of 2500 Hz on the counter, symmetrical trace on the gray on the drum, a photo for example, will read somewhere between 2500 and 2000 Hz. Good gray scale can be achieved if P2 is adjusted properly. Disconnect all test equipment. Turn the service switch on. Turn the output level control, R4, to a minimum. Turn on the transmitter and adjust the deviation by advancing the output level control. Keep in mind that F5 emission may only be as wide as a normal ssb signal on the low bands and only as wide as a standard a-m signal on 2 meters. That is, on 2-meter fm, keep the deviation at 6 kHz or less.

operation

The sending machine has the material to be sent while the receiving machine has the receiving blank installed, both with the overlap of the paper over the red line

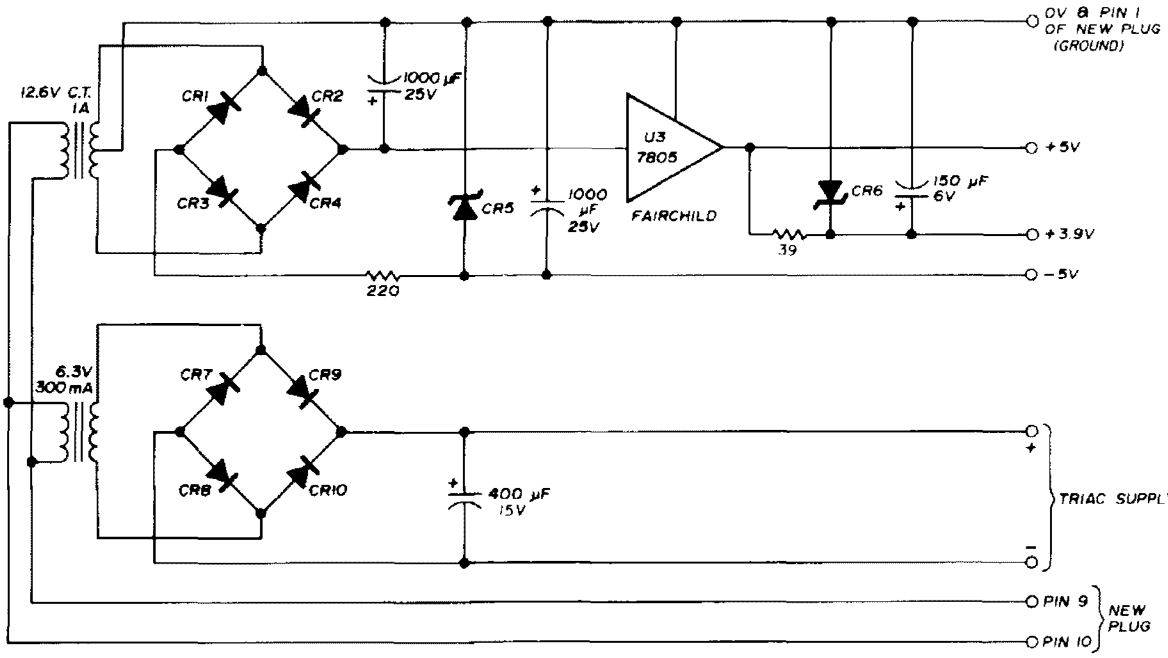


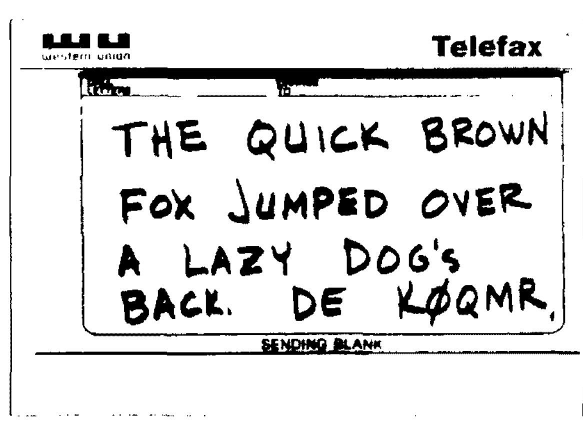
fig. 7. Power supply.

and the trailing lip of the paper located so that the stylus drags over the paper. Both machines are started, the OUTGOING button is pushed on the transmitting machine, and the INCOMING button is pushed on the receiving machine. Several seconds later the neon lamp will light. This indicates the tubes are warm and the machine is ready to use. Wait several seconds after the light is lit to be sure the receiving machine is ready and that sync has occurred. You can tell if the receiving machine is in sync when the hesitation in the gray-colored motor has stopped and the motor runs smoothly. The operator at the transmitting end pushes the drum feed button when he assumes the receiver is in sync (several seconds after the neon lamp has lit). After the button is pushed, the neon lamp will go out, signifying that a picture is being transmitted. The receiving machine will automatically detect the picture and cause the drum to feed. When the picture scan is complete, the machine will automatically stop, and the drum will return to its normal position.

If you push the wrong button, be sure to push the STOP button before pushing any other button. That is, if you push OUTGOING and meant to push INCOMING, push the STOP button before pushing the desired button. This is to get the logic straight again.

summary

We find it easy to just hook the setups back-to-back for demonstration purposes and testing, or a landline may be used if a radio link is not desired. If wired back to back use two wires and install a T-R switch. Simultaneous connection of output to input causes the drum to feed prior to sync acquisition. If hum level affects sync acquisition, adjust the receiver volume control to a lower setting or install a Butterworth filter between the receiver and the input to the limiter.



An actual transmission. Note the contrast between black and white and note the positive picture which was inverted electronically. The words "sending blank" were transmitted.

The Telefax machine conversion basically consisted of relay rewiring. It is still well suited for its original emission of negative pictures without using any of the new parts on the circuit boards. The basic sending and receiving functions of the machine are left undisturbed. The control system was rewired to cause the drum to feed horizontally with a pushbutton and for sync sending.

The St. Louis Amateur Teleprinter Society (SLATS) is a highly technical oriented group that prides itself in projects such as this. We find it quite handy to send schematics, photos, news clippings, etc., along with furthering our knowledge of linear and nonlinear circuits in specialized radio communications.

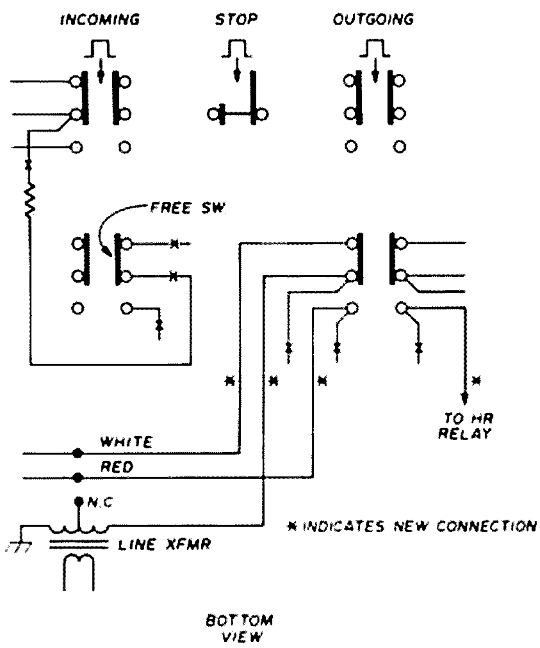
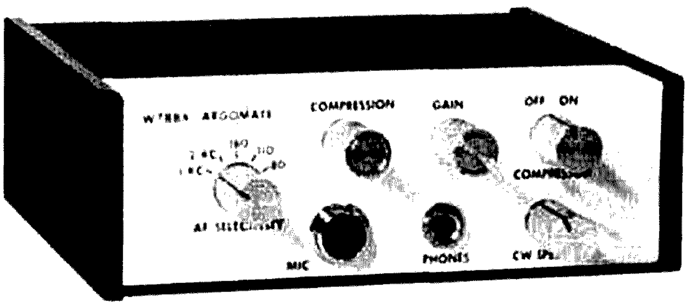


fig. 8. Incoming - outgoing switch modifications.

The author wishes to thank KØDOK, who sparked the development of the project; WAØIDS, who contributed the machine diagram and immense information on the machine; and the SLATS members, who inspired the article and acted as guinea pigs to prove the article could be understood.

A printed circuit board is available from the author for $6.50 postpaid or a full kit of parts including the circuit board for $62.00 postpaid in the U.S.A.

ham radio



introducing the argomate

Speech processor, ssb and CW filters, and keyer — all are combined in this desk-top companion to the Ten-Tec Argonaut transceiver

Howard F. Batie, W7BBX, Falls Church, Virginia 22042

The Ten-Tec Argonaut leaves little to be desired by the QRPP enthusiast. However, in the 5-watt and under area, anything that will improve performance is well worthwhile. This article describes a tailor-made companion for the Argonaut transceiver, but individual modules may be used with transmitter-receiver combinations as desired. Project objectives were:

1. Speech compressor with variable microphone gain and compression.

2. A 2-kHz ssb splatter filter.

3. CW filter, vary narrow passband, no insertion loss.

4. Keyer with self-completing characters, 10-60 wpm.

5. Dc power supply, 12-14 volts.

6. Minimum current drain for portable operation.

7. Detachable key for ssb-only or mobile use.

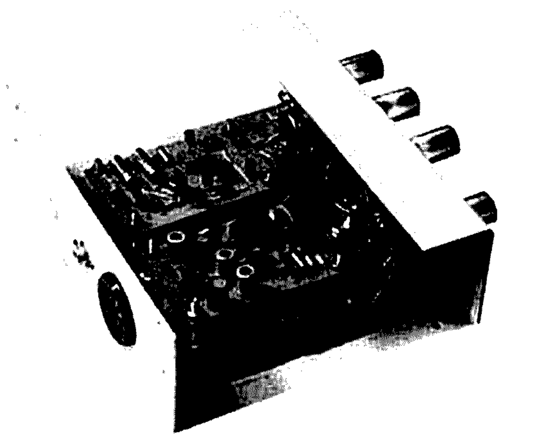
8. Small size and weight.

the argomate

A review of the current literature showed that no better audio filters could be found than the MFJ Enterprises lowpass and CW units, which are extremely small, rugged, and require only a few milliamps of current. Recent editions of the ARRL Handbook[1] contain an excellent speech processor circuit, which provides up to about 3 dB processing gain. For its cost and simplicity, the IC keyer by W7ZOI[2] was a natural choice. The finished product is a versatile adjunct to any station, high or low power, fixed or mobile.

A block diagram showing the interface of the two units appears in **fig. 1**. The ssb and audio filters provide optimum performance while retaining the agc feature on the filtered af signal only.

Minor modifications required to the Argonaut to accept the filter circuits are described; other receivers should be modified similarly after carefully checking the receiver schematic. If audio filters are outboarded from the speaker or phone jack less satisfactory audio filtering and agc characteristics will result. No mods are necessary to the transmitter for the speech processor other than adjusting transmitter drive and mike gain/compression levels for optimum clarity and performance.



Construction of the Argomate. Compressor board is on the left-hand side, keyer board is beside it. CW and lowpass filter boards are mounted under the compressor and keyer boards.

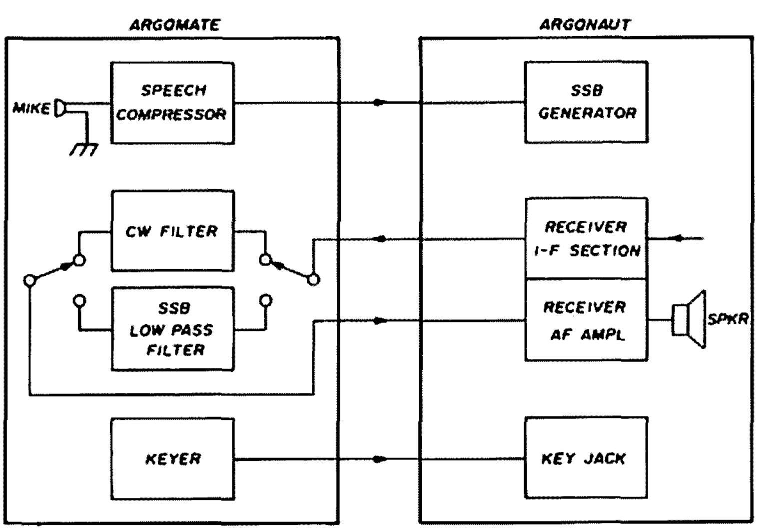


fig. 1. Interface between the Argomate and the Ten-Tec Argonaut transceiver.

keyer

The W7ZOI keyer (**fig. 2**) can be easily constructed on a 2 x 3-inch PC board. It uses a μA747, which is a pair of μA741Cs in a 10-pin TO-5 package. Resistor R2 is used to adjust the relative length of the first two dits to provide even spacing. The dot-dash ratio is determined by C3, C4; C4 is used only for the dot, and both C3 and C4 are used in parallel for the dash. Q4 collector provides for keying a positive voltage to ground (20V or less). The keying transistor will handle up to 50 mA without a heatsink.

compressor

The speech compressor details, operation, and adjustment are well treated in reference 1. Coil L1 can be either a UTC DO-T8 or UTC ML-6. The circuit is shown in **fig. 3**.

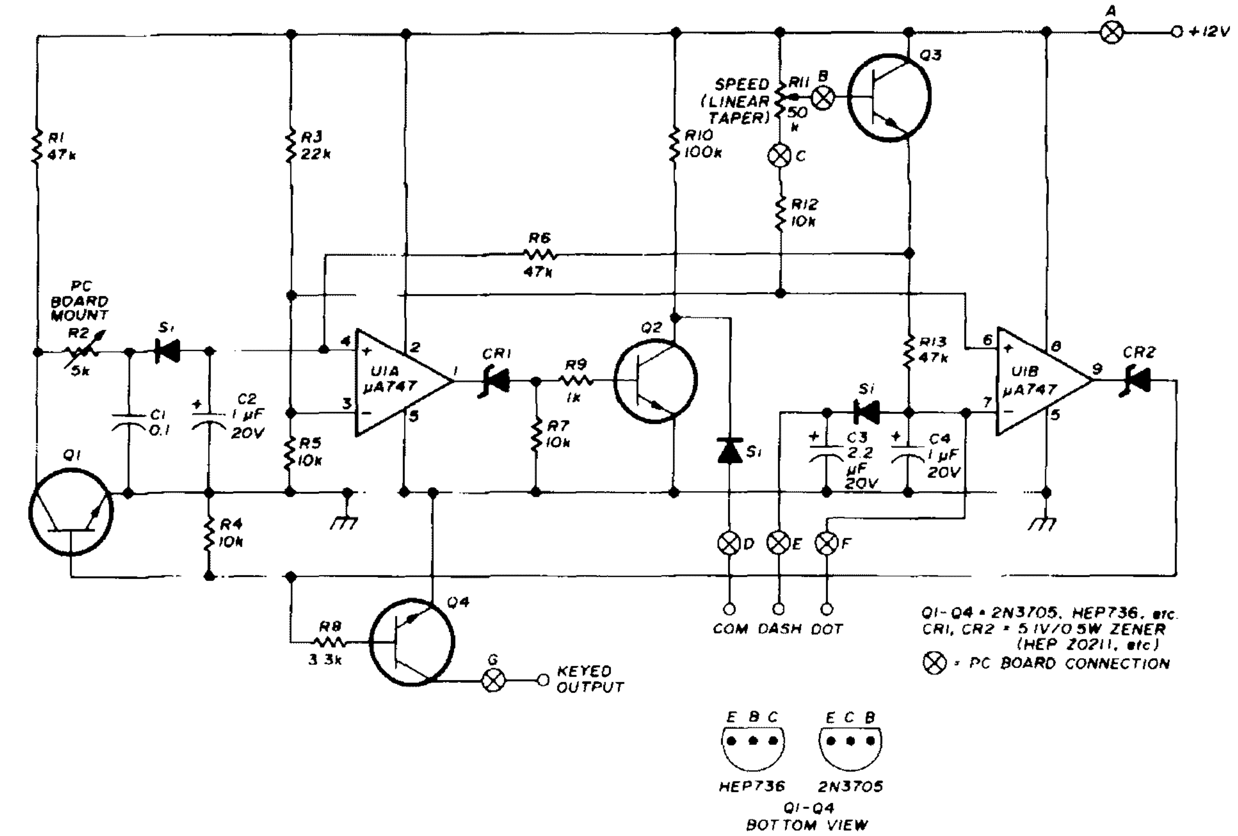


fig. 2. Argomate keyer schematic. Original circuit design, by W7ZOI, is described in the ARRL Handbook (reference 1).

cw filter

The CWF-2 filter is made by MFJ Enterprises and is available in either kit form or pre-wired and tested. Af selectivities of 180, 110 and 80 Hz are provided. This truly remarkable performer is a must for any CW operator. In the 80 Hz position, rolloff is 60 dB per octave, virtually eliminating any QRM not zero beat. Insertion gain (up to 2.4) is present in all positions. The high input impedance (680k) and low output impedance (less than 2 ohms) eliminate the

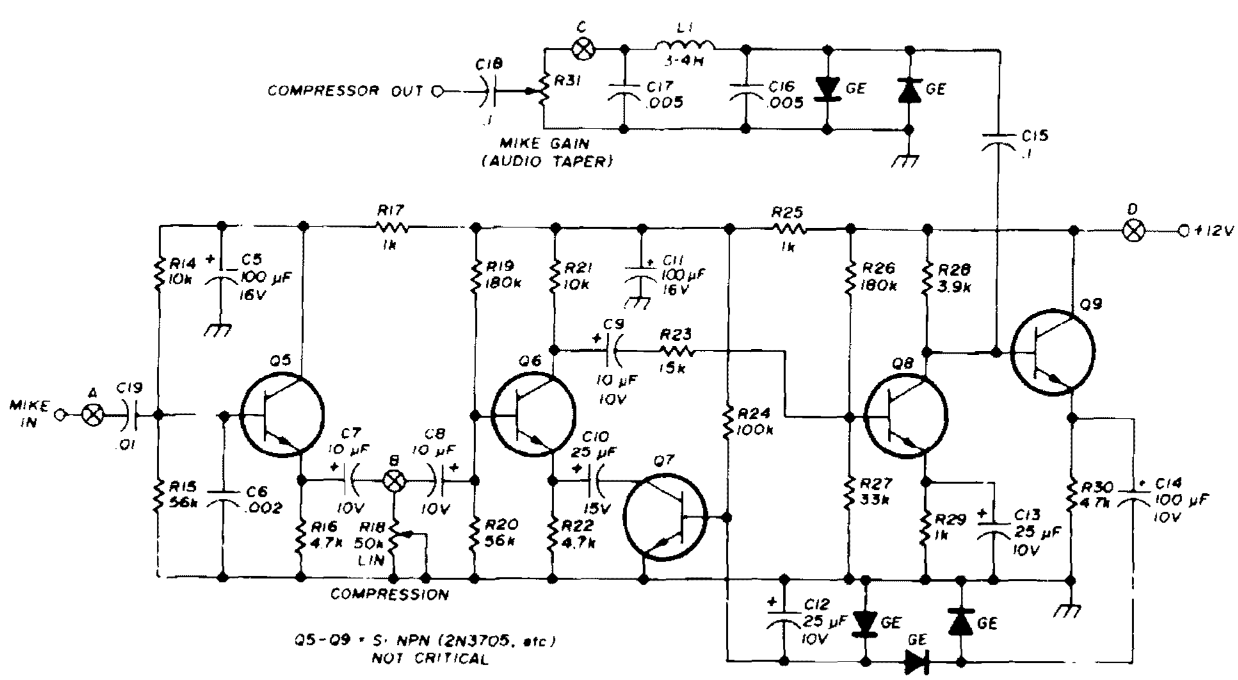


fig. 3. Speech processor.

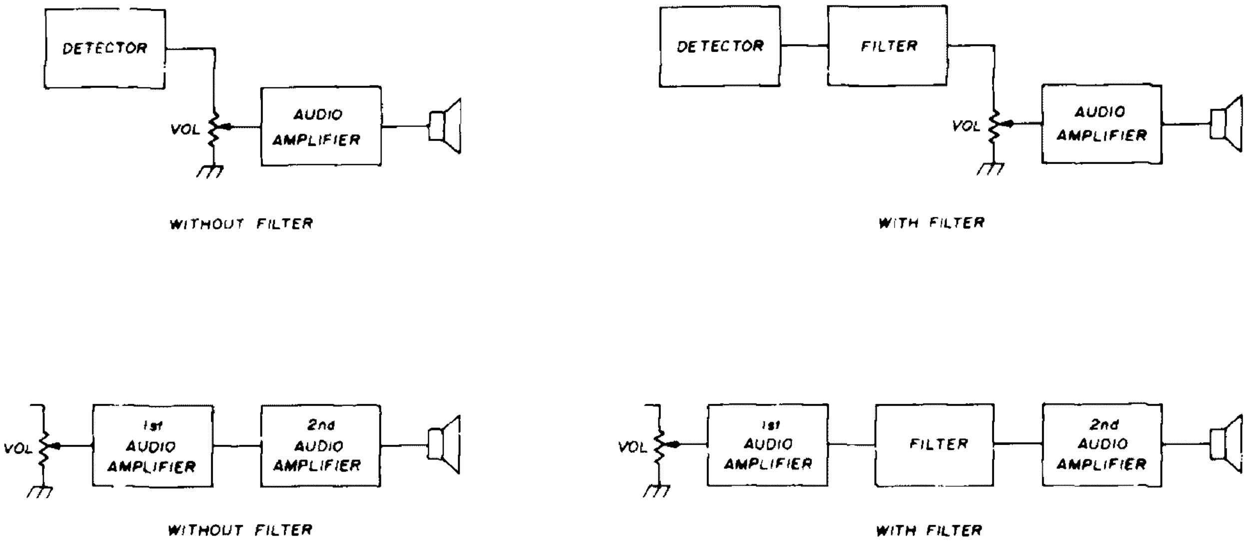


fig. 4. Method of connecting the CW and ssb filters. Better performance results if filters are inserted ahead of the audio amplifier.

necessity for receiver i-f/audio impedance modifications. The unit can be used with a 6-30 volt supply and draws 2-8 mA. The low-Q circuit design eliminates ringing, which is common in most conventional active filters. A 2 x 3-inch PC board is provided by MFJ.

low-pass filter

The LPF-1 low-pass audio filter is similarly constructed on a 2 x 3-inch PC board. This filter is used only in the receiver audio line to reduce ssb splatter above about 2 kHz. The cutoff frequency can be adjusted as desired by changing the values of eight resistors on the PC board. Received audio intelligibility of a male voice is not impaired as long as the cutoff frequency is above about 1.5 kHz.

audio filter insertion

Although both the CWF-2 and LPF-1

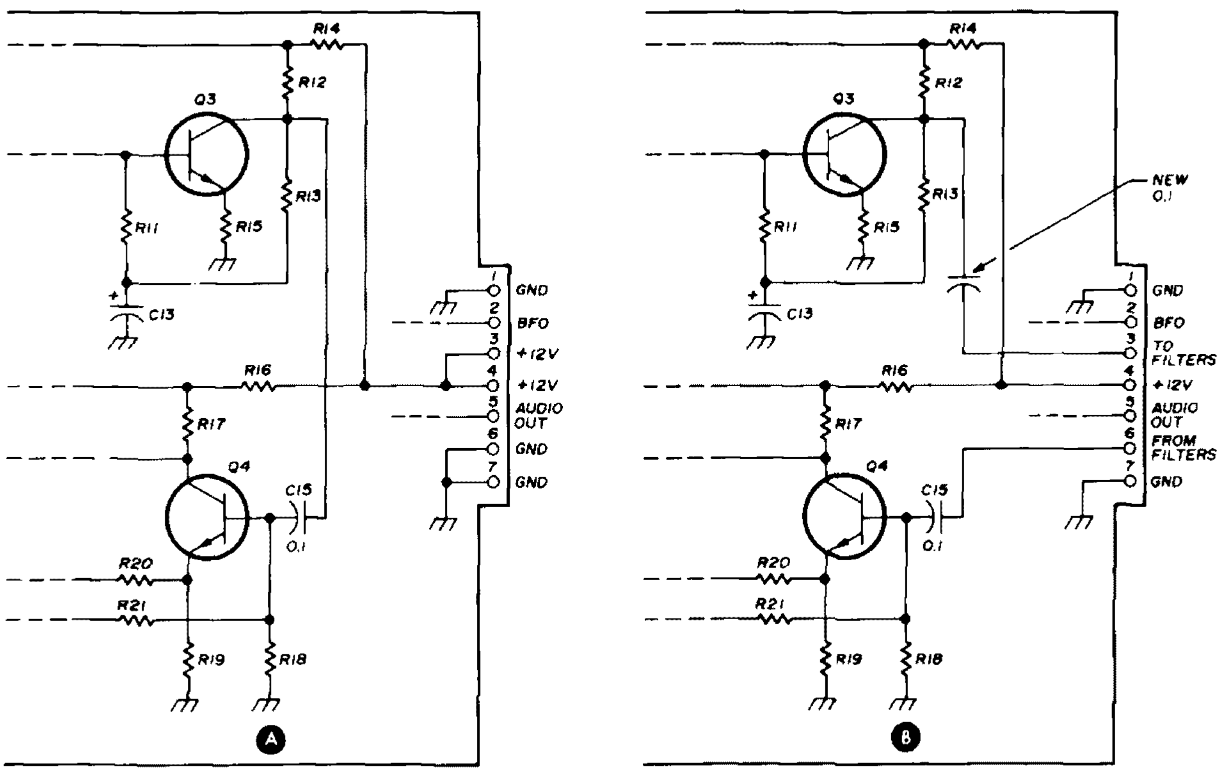
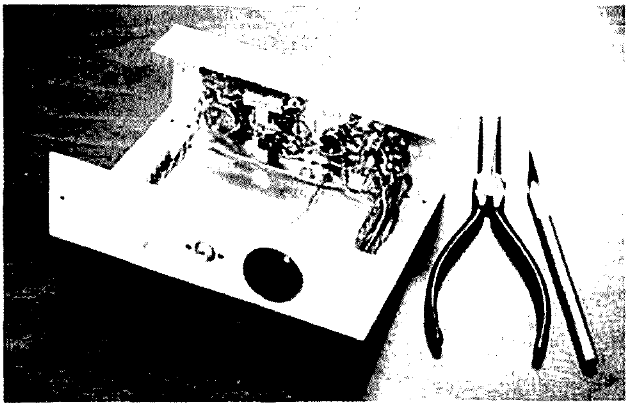


fig. 5. Argonaut i-f board modifications. A and B show before and after modifications. Component numbers are those of the Argonaut.

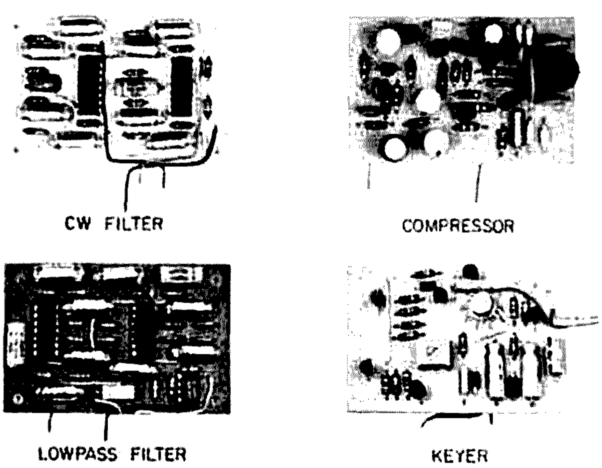


Front panel switching and control wiring.

filters can be connected directly to the receiver output jack, far better performance will be realized if the filters are inserted prior to the final audio amplifier in the receiver. Fig 4 shows recommended connections.

argonaut i-f board modification

Fig. 5 shows the modifications to the Argonaut i-f board for insertion of the filters. Mini-coax, such as RG-174/U, should be run to the rear panel connector for the filters. To perform the modifications, the copper foil joining pins 3-4 and 6-7 must be broken as well as the foil connection from C15 to the collector of Q3. C15 and the new 0.1 μF capacitor may be soldered to the top side of the i-f board if necessary. Unsolder the wire connected to pin 6 of the front terminal pin jack for the i-f board and resolder it to pin 7 (ground). The Mini-coax connections are then made to the i-f board



Four circuit boards used in the Argomate.

terminals 3 and 6 and run to the rear panel accessory socket.

argomate chassis construction

A Ten-Tec JW-7 cabinet provides a handsome matching enclosure for the Argomate. Additional side shielding of aluminum sheet was added to prevent stray hum and rf pickup.

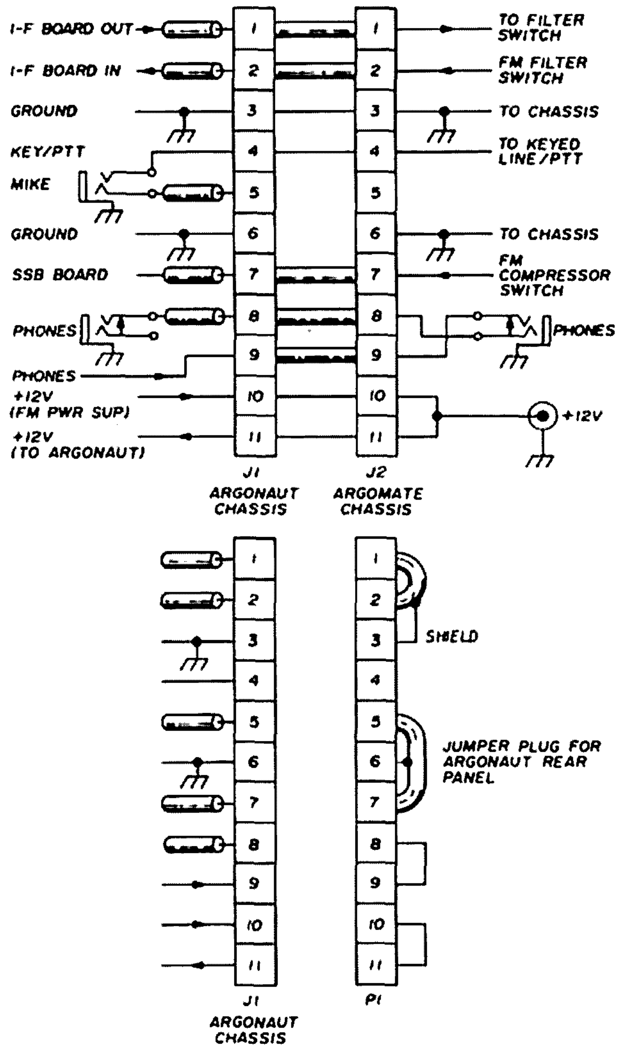


fig. 6. Interconnection wiring between Argomate and Argonaut chassis.

The four 2 x 3-inch PC boards were mounted in two stacks of two boards each in the horizontal position, with the CWF-2 and LPF-1 on the bottom. An 11-pin connector provides cable connections to the Argonaut, including power for the Argomate, as shown in fig. 6. A three-wire connector is provided for the cable to the Ten-Tec KR-1 paddle assembly. The mike PTT and keyer output

lines are connected together, as well as to the key jack on the Argonaut rear panel. Check the individual keying/PTT circuit before using this arrangement with other units.

The four PC boards are interconnected to multiple-pin terminal strips rather than with direct point-to-point wiring, which makes a neater installation and allows one or more PC boards to be removed if necessary. The terminal lugs serve as handy test points for every PC board input and output. **Fig. 7** illustrates the chassis connections. A jumper in the +12-volt leads on TS3 (pins 11 to 12) permits measuring the current drawn by each PC board when a milliammeter is connected between the two pins. **Fig. 8** shows switch and interconnection wiring.

The chassis cabinet and additional aluminum sides provide adequate shielding for all wiring and components inside the chassis. However, in areas where ac hum or rf pickup are severe, additional RG-174/U will reduce these effects inside the cabinet. Coax or other shielded connectors must be used between the Argonaut and the Argomate.

An additional 11-pin jumper plug must be used when operating the Argonaut without the Argomate. This plug rewires the i-f/audio connections to the original condition. The Argonaut +12 V supply line is jumpered in the plug and the Argomate rear apron plug. This feature prevents unauthorized operation if no plug is inserted into the Argonaut but may be eliminated if desired.

To install the 11-pin jack, J1, on the rear panel of the Argonaut, it will be necessary to relocate the adhesive serial number and the existing rf output jack.

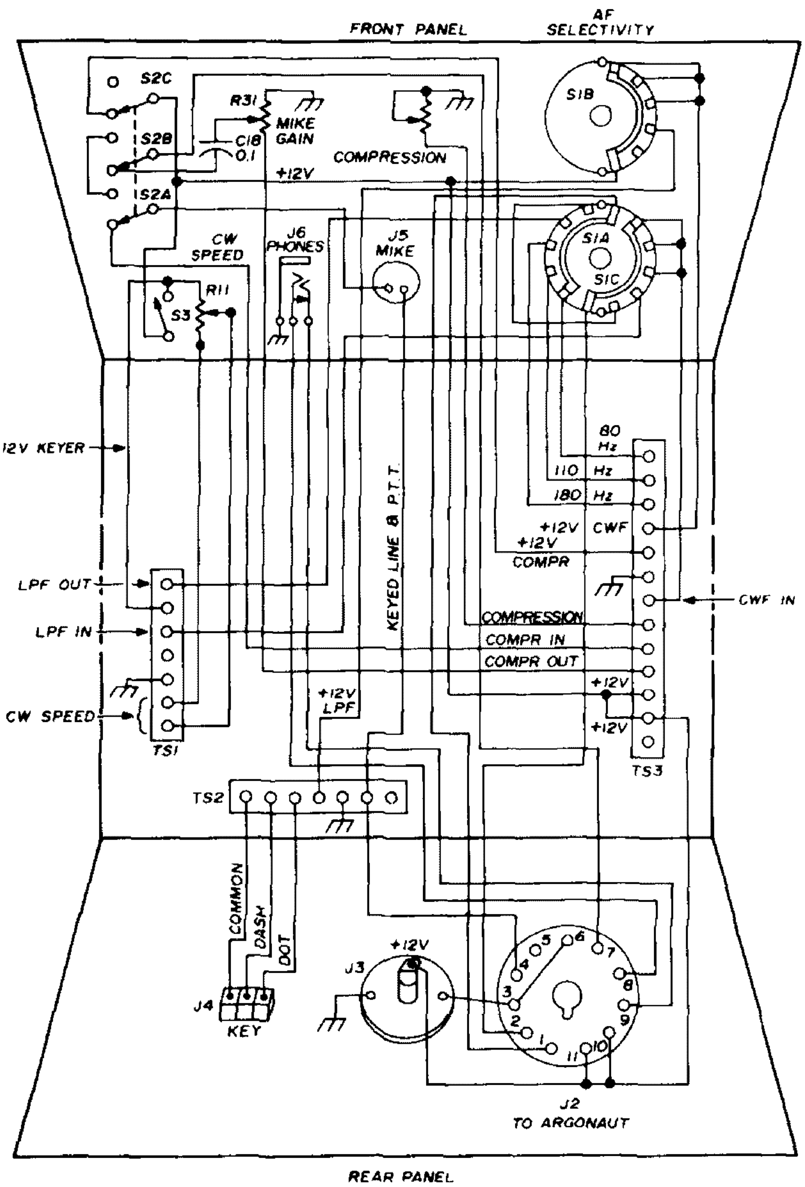


fig. 7. Pictorial representation of chassis connections.

Space is limited, but with judicious placement, sufficient clearance for the relocated rf output jack and J1 is available. A resistor and a capacitor are soldered to the Argonaut microphone jack; relocate these with the mike lead to pin 5 of J1.

operation

For ssb, the mike connector is brought out to the Argomate front panel for

convenience and easier accessibility. The Argonaut mike connector is not used when operating the Argomate; however, the speech compressor may be bypassed on the Argomate front panel by S1. The Argonaut mike jack is reconnected to pin 5 of J1 only. The Argonaut mike PTT line remains connected to the keying line and is additionally connected to pin 4 of

Use of the Argomate will not require modifications to an external linear amplifier as long as the linear can accept the increased duty cycle when operating the ssb speech compressor. The Argomate mike gain and Argonaut ssb drive controls can be easily adjusted for the most effective speech compression and intelligibility.

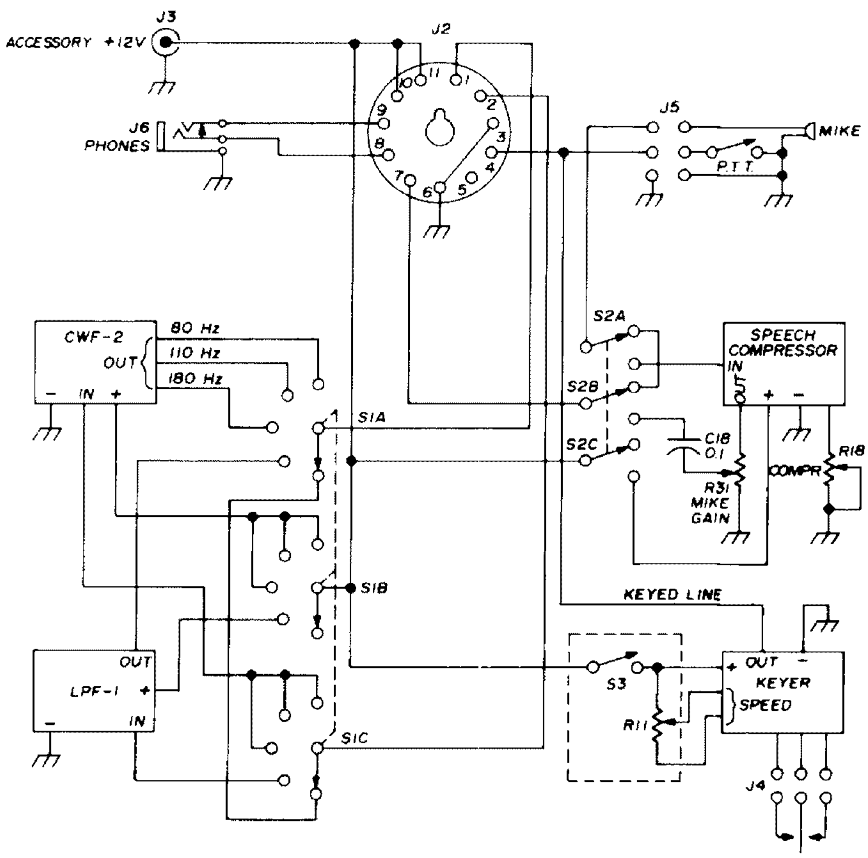


fig. 8. Details of component interconnections.

J1. For ssb operation without the Argomate, the Argonaut mike connector is activated by the jumper wire in plug P1.

On CW, full QSK operation is retained with or without the Argomate. The Argonaut key jack and ssb PTT controls remain activated when the Argomate is connected.

The Argonaut and Argomate phone jacks are connected in series; therefore, phones or an external speaker may be connected directly to either jack, which disables the Argonaut internal speaker. With no phones or speaker connected externally to either the Argonaut or Argomate, the Argonaut internal speaker remains connected.

use with separate transmitter and receiver

For use with a separate transmitter and receiver, the Argomate can easily be adapted with cables to each by addition of a separate multi-connector plug. Since the speech processor, keyer, and audio filters can be controlled independently, only those functions desired need be included in the Argomate. However, the design shown here has the advantage of requiring only a single +12-volt power supply for all functions, while providing the most essential improvements on both transmission and reception for the most discriminating ssb and CW operator.

mobile operation

The Argomate can tolerate supply voltages to about +18 volts; however, if the Argonaut/Argomate combination is used for mobile work, care must be taken to ensure that the alternator or regulator output of the car's electrical system is not higher than the recommended Argonaut supply voltage, +14 V. To provide proper voltage regulation when a voltage dropper is needed, the voltage drop must be independent of the difference between the current drawn on transmit and receive. This implies that the voltage dropper must have near-zero ohms internal impedance. This is easily accomplished by using a series of silicon diodes from the car battery (or cigar lighter) to the Argonaut +12 V supply input, as shown in **fig. 9**. To determine the number of diodes required, measure the battery voltage with the engine running, and figure on 0.6 volt drop across each diode (0.3 if germanium diodes are used). Each diode should be rated for at least three times the maximum current drawn by the complete installation in the transmit mode.

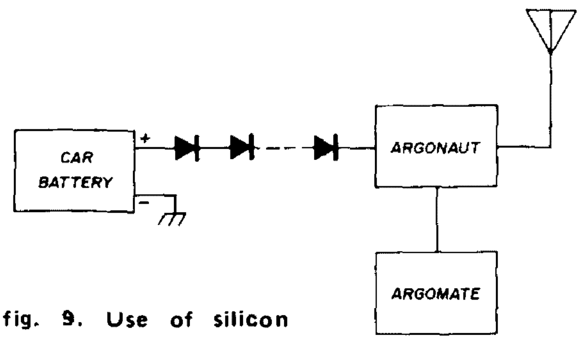


fig. 9. Use of silicon diodes between car battery and Argomate/Argonaut for mobile operation (see text).

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2. Wes Hayward, W7ZO1, "An Integrated Circuit QRP Keyer," *QST,* November, 1971, page 38.

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ham radio

low-cost monitor receivers for two-meter fm

How to convert fm broadcast receivers to 144-MHz fm service

James E. Trulove, WB5EMI, 1409 SW 70th, Oklahoma City, Oklahoma

If you are an avid fm repeater user, you have probably had occasion to wish you could monitor the frequency at your leisure, without having to blow your bankroll on an expensive vhf monitor receiver. This article proposes an alternative: converting an fm broadcast band receiver to the vhf high-band. If you have a little-used a-m/fm portable or even an old clock radio lying around the house, chances are that it can be converted to two meters. I have performed such surgery on several fm radios with surprising success. An additional benefit is being able to monitor police and public service broadcasts as well. The basic modification is quite simple and straightforward and should be easy to complete on most fm radios.

The superheterodyne circuit used in fm broadcast radios primarily uses two tuned elements to make the receiver tune a particular band. Those elements are the LC tank circuits in the rf amplifier/preselector and in the local oscillator. This means that all you must do to change the received frequency is to vary the range of the resonant frequency of the preselector and the local oscillator tanks. This can be done easily if, the radio uses transistors that will operate well at the higher frequency, and the radio has sufficient sensitivity at the higher frequency to enable you to hear the vhf transmissions. It is also very important to have readable schematics, as this helps considerably in making the proper modifications.

A good rule-of-thumb guide for selecting a receiver to modify is to find a recent receiver model (implying better high fre-

quency transistors) with excellent sensitivity. A good way to judge sensitivity is to retract the antenna all the way — or disconnect it altogether — then try to tune in an fm station normally. If you still receive most stations at full quieting, then the receiver should perform well with the weaker signals present in the public service and amateur bands. However, don't expect the superior sensitivity and selectivity of a $150 crystal controlled, dual-conversion receiver.

converter (which serves the dual purpose of local oscillator and mixer). The rf amplifier is originally designed to tune 88 to 108 MHz, tracking with the local-oscillator tuning. This must be changed to cover the range of 140 to 160 MHz, providing a central frequency of about 150 MHz. You can achieve this by decreasing either the capacitance or the inductance of the rf tank. However,

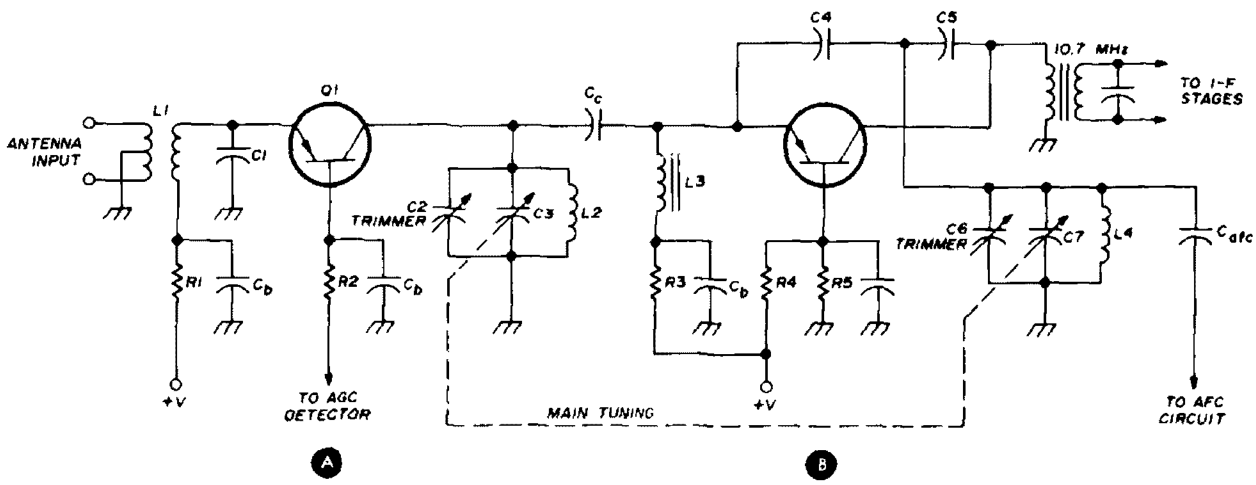


fig. 1. Typical rf amplifier and converter stages found in modern fm broadcast radios. In the rf amplifier (A), L1 and C1 form the low-Q input tuned circuit and L2, C2 and C3 make up the high-Q output circuit. In the converter stage (B), local oscillator tuning is controlled by C6, C7 and L4.

the modifications

Most of the fm radios currently in use, use some variation of the rf amplifier and converter circuits shown in figs. 1A and 1B, respectively. These two circuits normally precede the a-m section of the receiver. Bandswitching disables the fm section and changes the role of the first two transistors in the a-m section from i-f amplifiers to an a-m band rf amplifier and converter, respectively. The modifications, then, will be directed at the fm portions of the radio only, leaving the a-m section substantially unchanged.

The primary modification is to the inductances in the rf amplifier output tank and in the oscillator tank of the

reducing capacitance will also have the undesirable effect of reducing the tuning range since the main tuning capacitor is not changed. Thus the inductance must be lowered. This will expand the tuning range somewhat, but will not seriously limit receiver performance.

A quick calculation from the resonant LC formula shows that to increase the resonant frequency by a factor of 1.5 you must decrease the inductance by a factor of roughly 0.5. Of course, this could be done by removing the coil presently installed and replacing it with one of half the inductance. To make matters simple, all you have to do is duplicate the coil and place it in parallel with the present one.

Take a look in the radio you have: the coils should be about three or four turns,

either air wound or on a plastic coil form about 3/16-inch (4.8 mm) in diameter. You may need to do a little visual circuit tracing to find the exact coils involved. The precise dimensions of the new coil are not too critical, as the inductance can be changed quite a bit by compressing or

fig. 2. Layout of the main tuning capacitor. Preselector and local-oscillator trimmers for the fm section are marked on top. An unused connection to the high side of the preselector trimmer is at (A); the corresponding connection for the LO trimmer is at (B). A ground strap for the modifying coils is available at (C).

expanding the turns of wire. Three or so turns of number-20 bare wire was found to work well on various radios.

The same technique is used to raise the local oscillator frequency. It doesn't go up by quite the same factor, so you might want to use a half turn or so less for this coil.

The next thing to decide is where to place the coil. Shown in fig. 2 is the tuning capacitor module commonly found in these a-m/fm radios. It has four sections, two of each rf and oscillator tuning for the a-m and fm sections. The four trimmer capacitors in the top of the case each adjust one section of the main ganged capacitor. As seen in fig. 1, one side of each capacitor is connected to ground. As both fm trimmers have connections appearing on the top of the case, this is probably the easiest place to mount the two new inductors. Placing them across the trimmers to ground amounts to paralleling them with the original inductors. The pictorial in fig. **3** shows the placement of the new inductors. Note that they are placed so that the axes are 90° apart. This minimizes coupling between the two stages.

One word of caution: some receivers are designed so that the unused lead of the trimmer capacitor is not at dc ground. If you have such a receiver, install a 0.001 μF ceramic disk dc blocking capacitor in series with the new coil to prevent damage to the radio. Better yet, choose a capacitor with leads longer than 2½ inches (64 mm) and form the coil from one of the two leads.

alignment

Rough alignment is really fairly simple, and it will probably be quite sufficient if you only wish to tune one small band of frequencies, such as 146 to 148 MHz. First you need a fairly strong signal at the frequency you are trying to tune. A dummy-loaded transceiver should suffice for setting up the two-meter band, as most dummy loads have a little rf leakage. If this method is not convenient, you may be able to use transmissions on radiotelephone at about 150 MHz, or even your local repeater.

With the signal source turned on, adjust the main tuning dial to a convenient location near the middle of its tuning range. Then, with a small screwdriver, carefully turn the local oscillator trimmer until the signal source is well tuned in. If you can't find the signal, try adjusting the inductor by compressing or expanding the windings. Then repeat the above procedure again. You may need to rewind the coil if you still have difficulty. If you have a variable signal generator or a grid-dip oscillator, use it to find where you are tuned. Remember that you

haven't adjusted the preselector yet, so you should be receiving two frequencies: the desired one and its image. The local oscillator will be between these two frequencies.

Assuming, then, this alignment has been accomplished, the preselector must be set up. This is easy. Leaving the signal source on and the radio tuned in to it, adjust the preselector trimmer for maximum signal. Since the tuning of this tank circuit is coupled through the converter transistor to the local oscillator tank, there will be some interaction, so go back and retouch the oscillator trimmer. Repeat this procedure a few times. If you have difficulty with the preselector, use the hints given in the previous oscillator alignment. A grease pencil may be used to make reference marks on the receiver tuning indicator for the new band.

This takes care of the initial alignment. You should now be able to tune in stations well on the band of frequencies around the one where the alignment was performed. But proceed with caution; the following "fine tuning" adjustments are much more difficult to make and may well require some specialized equipment. If you are satisfied with the sensitivity, selectivity and tuning range of your converted radio, you are finished with alignment. For the more adventurous souls, or the perfectionist, read on!

input tuned circuit

The antenna input is normally a balun transformer with a tuned secondary as shown in **fig. 1**. The Q of this tank circuit is low, broadbanding its response, and it primarily provides protection from frequencies considerably removed from the fm band. This circuit can best be retuned by removing the capacitor and substituting one of half its value.

tracking

If you wish to tune the entire public service band, proper preselector tracking needs to be taken into account. Tracking refers to the fact that for proper tuning the preselector must be tuned exactly 10.7 MHz away from the local oscillator frequency. When originally manufactured, your fm radio was adjusted to do just that.

Normally, the receiver is designed to have three points of perfect tracking, shown as zero tuning error in **fig. 4**.

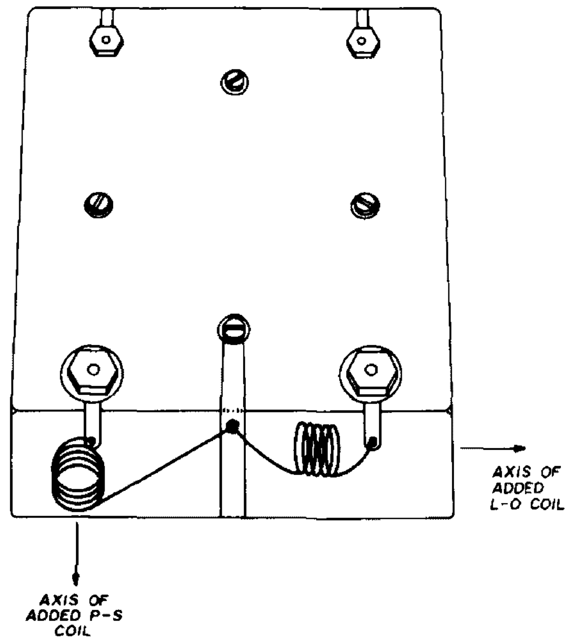


fig. 3. Mounting the new coils. The coils should be soldered between the unused trimmer tabs and the ground strap with their axis at right angles.

Tracking varies over the tuning range because of the interaction of the two steps.

Since several of the component values have been changed (as well as the operating frequency), you can presume that the preselector no longer tracks the received frequency. The solution is to adjust the value of the preselector inductance, then readjust the preselector trimmer capacitor until the receiver more or less tracks in the range of interest.

Pick two points on the dial to adjust the tracking, one at the high end and one at the low end. If you have access to a signal generator, any alignment points will do. If not, set the tracking at two discrete frequencies to which you wish to listen. Alternately adjust the preselector inductance (by compressing or expanding the coil turns) and the preselector trim-

mer capacitor at both high and low ends of the dial until the preselector stage tunes properly at both points.

i-f selectivity

As the commercial fm broadcaster uses 75-kHz deviation, broadcast receivers use an i-f bandwidth of about 150 to 200 kHz. Likewise, the detector is broadband. There are two methods of achieving this broadband response in the i-f stages: stagger tuning and resistive loading of the tuned circuit.

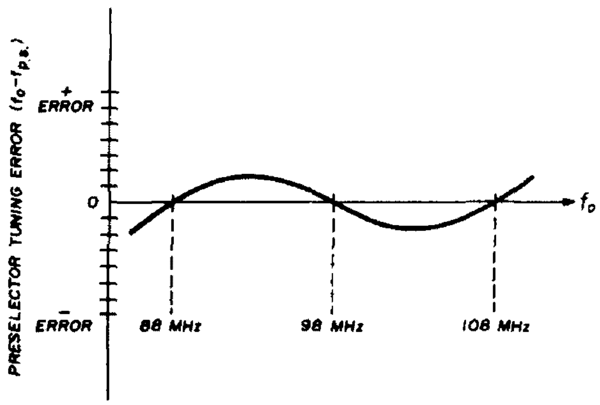


fig. 4. Points of perfect tracking, showing the three points of zero tuning error, where the tuned frequency (determined by the sum of the 10.7-MHz i-f plus the oscillator frequency) equals the preselector center frequency.

In stagger tuning, stages are purposely misaligned to yield an overall wide passband. This technique naturally decreases the total gain of the i-f system. Here the selectivity can be improved by tuning the i-f stages to the same frequency while receiving a fairly weak station or other fm-modulated signal source. Take care that the increased gain produced by your retuning does not cause oscillation, even if it means detuning one or more stages.

With resistive loading, a resistor is normally placed in parallel with the inductor, although sometimes the effective resistance may be in the coil itself in the form of small wire size and a large number of turns, or in the loading effect of the transistor. The only modification that is practical here is to clip out the discrete loading resistor. This should be done only if the resistor is not part of the *transistor* bias circuit.

Frequently a combination of stagger tuning and resistive loading is used. It is likely that retuning the stages will be sufficient.

automatic frequency control

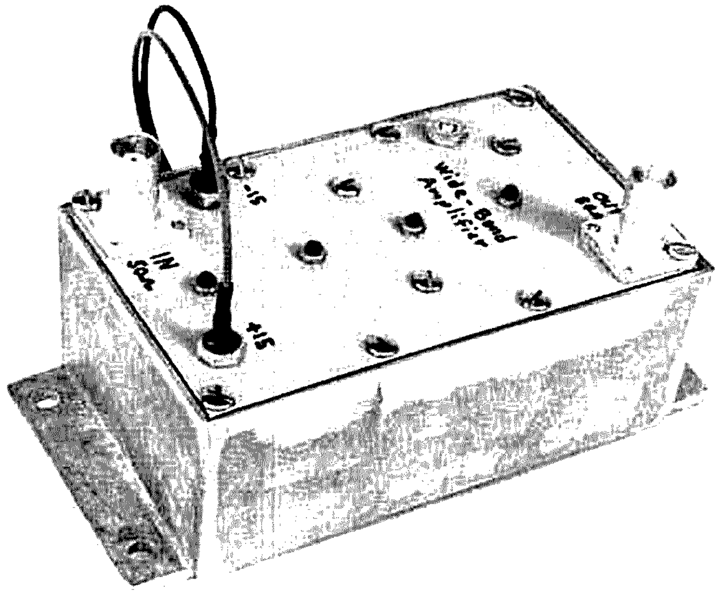
For some reason most fm radios are equipped with afc. This was great in the days of thermionic emission devices (tubes) where the source of instability was extremes of temperature as the filaments heated up. In transistor equipment, the afc circuit is more of an advertising point, and frequently introduces hysteresis, making tuning more difficult. While this is no real disadvantage for 200-kHz wide fm broadcasts, it is definitely harmful when trying to tune intermittent transmissions only 10- to 30-kHz wide. In the first radio that I modified, a poorly designed afc system actually pulled the local oscillator off frequency after each transmission began! The afc will also pull the receiver off a desired transmission onto a strong undesired one if it is within its range.

To disable this circuit, simply cut out the afc coupling capacitor, C_{afc}, shown in fig. **1B**. This capacitor couples the capacitance of the afc transistor, which varies with the magnitude of feedback, from the fm detector circuit. Clipping it out will only affect the tuning of the local oscillator tank and will not affect any dc components in the afc or detector circuit. The local oscillator should normally have sufficient stability to preclude serious drifting.

conclusion

It is hoped that this article has provided a simple way of monitoring two-meter transmissions. Converting a radio for a friend is a great way to give him a taste of two-meter operating. Performance should be adequate for casual monitoring, though selectivity and sensitivity will not be as ideal as in an expensive transceiver.

ham radio



wide-range
broadband amplifier

Distributed amplifier principles are put to use in a small package that provides 18 dB gain from 1 to 36 MHz

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

Broadband rf amplifiers are becoming quite common nowadays with the availability of ferrites and transistors with high gain-bandwidth products. The current fad is to spend all your design time on the ferrite transformers so that the 50-ohm output impedance can be transformed to some higher impedance that may serve as a reasonable collector or drain load for the transistor or fet you use. Such transformer design is beyond the ken of most hams, even if they could get the required ferrite cores. The general principles of broadband transformer design are described in two rather good articles.[1,2] Beyond these basic articles are a host of little practical tricks and facts, many of which are not available in text at all. One transformer manufacturer has gone to such lengths as to put a sheet of lead inside the epoxy case of his latest rf transformer when it was on loan to a large systems company so that the potential customer couldn't x-ray it to see how it was constructed (before quantity ordering).

distributed amplifier

The broadband amplifier described here does not use any special ferrite transformers, and uses plastic economy-type transistors. The construction techniques are easily within the ham-experimenter's ability, and the finished amplifier has rather good performance for its cost.

An old, neglected technique of broadband amplifier construction was selected

because it requires no special parts. The distributed amplifier is the name by which this amplifier goes, and it dates back to 1937. The distributed amplifier was originally designed to be used with vacuum tubes, so that the input capacitance of each grid could be lumped with the shunt C of an artificial transmission line on the input side of the amplifier. In a like manner, the output capacitance of each anode was lumped with the shunt C be used in the distributed amplifier. The base-emitter junction is used *in series to ground* with the shunt C of the artificial transmission line. In this way it is the *current through* the shunt C of the transmission line that drives the base of the transistor – which is just what the bipolar transistor requires. Reference 4, which is now over ten years old, describes a practical amplifier of this type as well as the basic design equations.

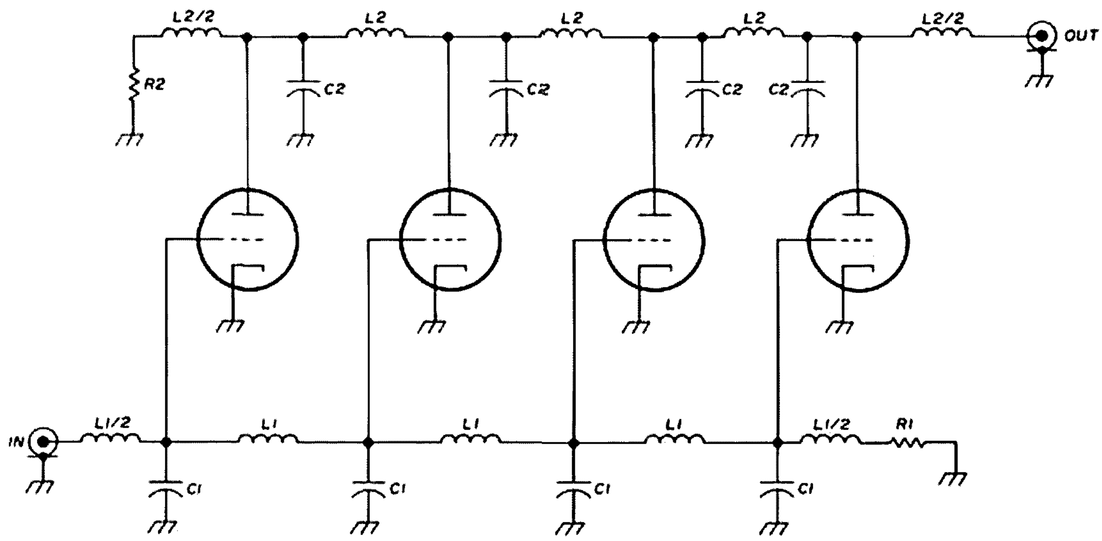


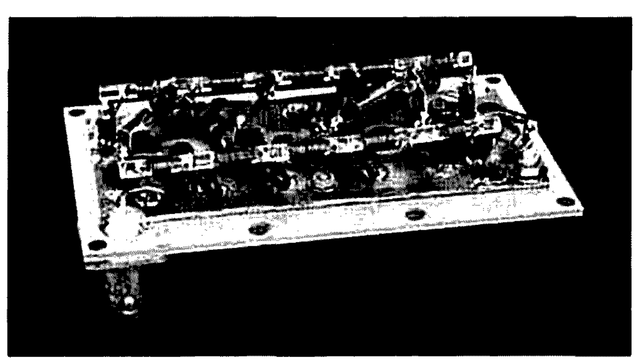
fig. 1. Basic distributed amplifier.

of an output transmission line. The basic distributed amplifier (with tubes) is shown in fig. 1. The gain contribution of each tube, in phase with the amplified wave as it passes down the artificial transmission line, adds to the contributions of the other tubes. Therefore, the whole distributed amplifier acts as if the transconductances of the tubes were all in parallel; i.e., the stage gains add.

bipolar transistor amplifier

You can easily see how the principles for a vacuum-tube distributed amplifier could be almost directly applied to one using fets. This has been done, in fact, in a technical article that will soon be published by Siliconix, one fet manufacturer.[3] It is a bit less obvious, however, to see how the principle of the distributed amplifier can be applied using bipolar transistors. Fig. 2 shows the method by which bipolar transistors can

The broadband amplifier shown in fig. 3 was built using the method described in reference 4. Since ±15 volts is a rather common type of power supply nowadays (because of the wide use of operational amplifiers), the amplifier was designed to operate on this source of power. The transistor type chosen for the amplifier was the Motorola MPS-U05, a silicon npn type in a small plastic power package.



Layout of the distributed amplifier showing the two delay lines. Input is to left, output to the right.

This transistor is truly a marvel of power handling and gain-bandwidth product, considering its cost is in the one-dollar vicinity. Fig. 4 shows how the gain bandwidth product and beta vary with current of the MPS-U05.

transistor, mica washer, aluminum washer and aluminum plate clamping. Silicone grease was used in all these thermal interfaces to increase heat conduction.

The inductors that make up the various L values in the artificial transmission

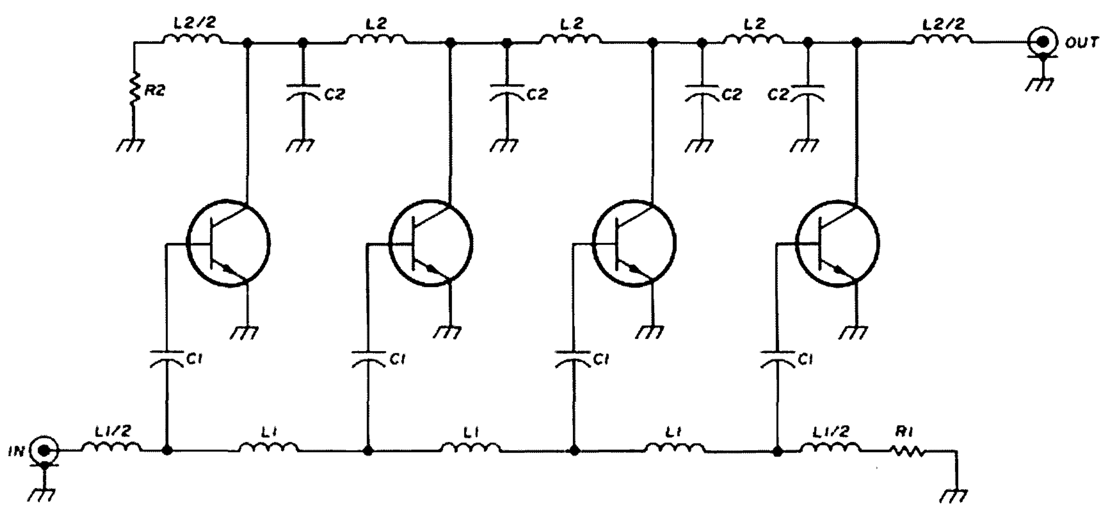
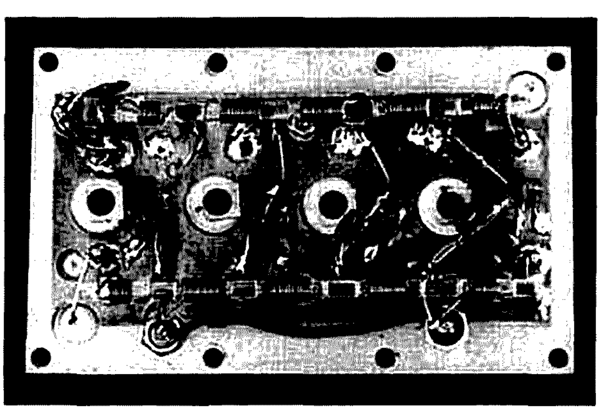


fig. 2. Distributed amplifier using bipolar transistors.

construction

When constructing the distributed amplifier, note that the four transistors must have their collector tabs installed in a heat sink. In this design the heat sinks are 1/2-inch (127 mm) diameter aluminum washers set into holes in the piece of copper laminate on which the circuit is built. The tabs of the transistors actually are mica insulated (electrically) from the aluminum washers, and the washers conduct the heat from the transistors to the aluminum outer plate of the box chassis. Nylon 4-40 screws were used for the



Plan view of the distributed amplifier showing the four transistor stages and home-made delay lines. Input is at right, output at left in this photograph.

lines were wound on two 1/8-inch (3.2 mm) diameter lucite rods (one for the input 50-ohm line and one for the output 50-ohm line). The dimensions of these multicoil structures are not at all critical. In fact, the first version of this amplifier was built using ten standard-value rf chokes and a handful of standoff terminals. The single rod, hand-wound coil structures were used simply to conserve space and save costs.

The 100-pF capacitors in the two transmission lines should be silver-mica types. The DM5 version of silver mica is smaller and easier to use here, but larger types should work. The emitter-bypass and output-coupling capacitors should be low-inductance ceramic types. The *Redcap* 0.1-μF capacitors by Erie are just fine. It would be better to use 0.01-μF ceramic disc capacitors (and have the low frequency end suffer a bit) than to use 0.1-μF capacitors of foil construction (mylars, etc.) if the low-inductance 0.1-μF types cannot be readily obtained. Do not parallel capacitors for emitter bypassing. Such practice seems like a good idea, but usually results in a vhf parallel-tuned circuit, which causes the amplifier to take off.

There are three 1000-pF standoff or feedthrough-type capacitors in the amplifier. These are used in noncritical points in the circuit for getting into the box chassis or where a tie point is needed. The 1000-pF standoff type is located at the junction of the two 300-ohm emitter resistors of the first two stages. A 1000-pF feedthrough type is located at the junction of the two 300-ohm emitter resistors of the last two stages and serves as the −15 volt connection to the outside of the box chassis. Between these two 1000-pF capacitors is a piece of hookup wire with a ferrite bead on it to damp out a possible vhf parasitic resonant circuit. These ferrite beads are in common amateur use now and are available from Amidon Associates.*

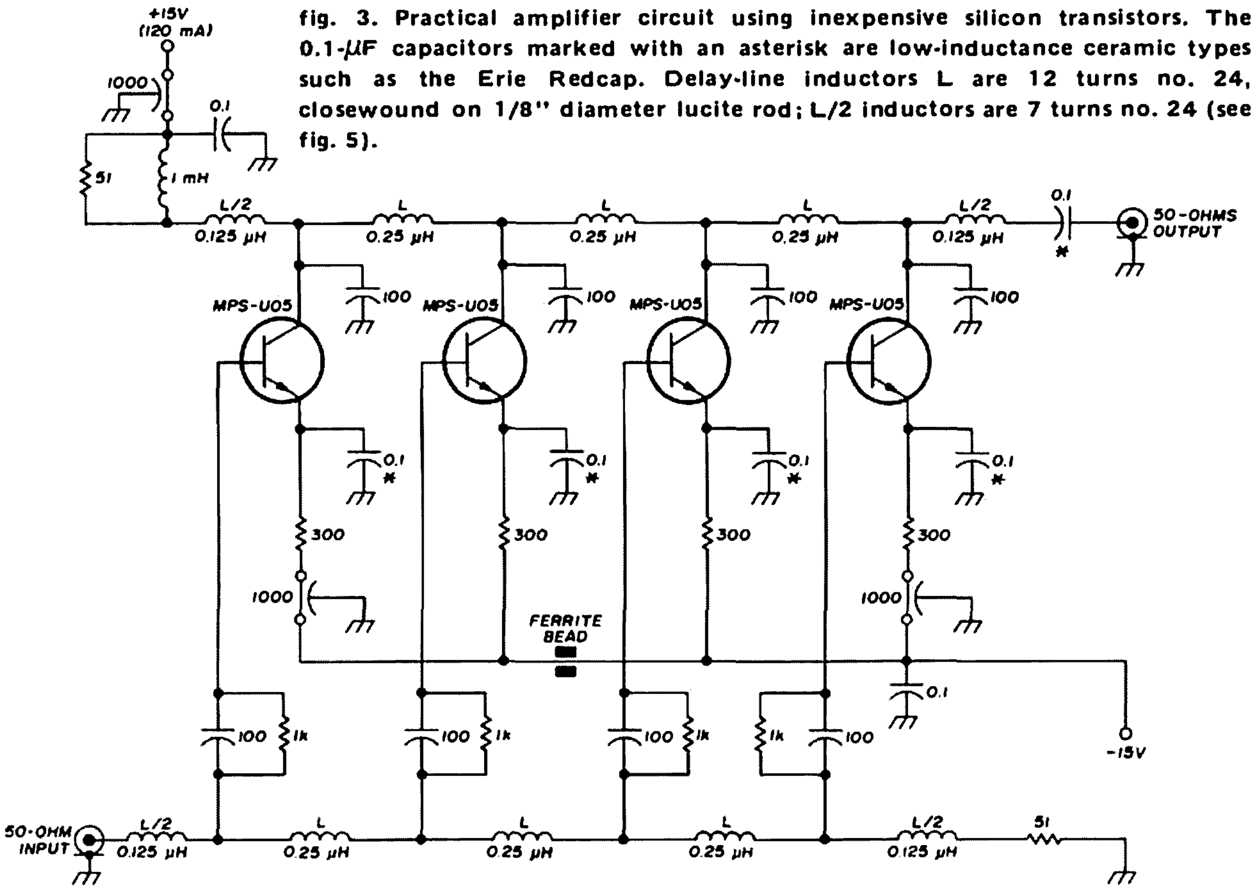
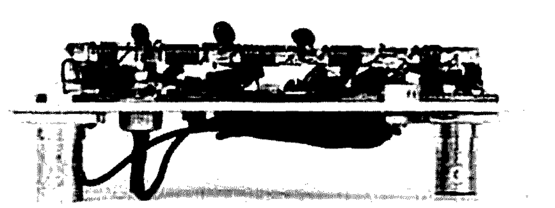


fig. 3. Practical amplifier circuit using inexpensive silicon transistors. The 0.1-µF capacitors marked with an asterisk are low-inductance ceramic types such as the Erie Redcap. Delay-line inductors L are 12 turns no. 24, closewound on 1/8" diameter lucite rod; L/2 inductors are 7 turns no. 24 (see fig. 5).

testing

With the input and output terminated in 50-ohm test equipment, the amplifier should be tested. The amplifier should draw about 120 to 150 mA with ±15 volts applied to it. The gain is about 18 dB, flat to within 3 dB from 1 to 36 MHz. The noise figure was measured only at 30 MHz with an A.I.L. automatic noise figure meter; it was 8 dB. The amplifier was capable of putting out +20 dBm, or 100 milliwatts before a compression of 1 dB was encountered. In fact, by increasing the 300-ohm emitter resistors to 1 watt and using ±30 volt supplies and higher-voltage MPS-U06 transistors, it was possible to get 1 watt output before 1 dB compression set in.



Side view of the distributed wideband amplifier showing position of delay lines.

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

applications

Assuming a careful job of construction and testing as described above, you now have a broadband hf amplifier. What can be done with it? The first thing that may come to mind is its use as a preamp for hf receivers from 160 through 10 meters. However, putting an 18-dB preamp ahead of most hf receivers will usually cause more problems than it will solve. The extra 18 dB will *reduce* the dynamic range of the total receiving system by nearly 18 dB, unless the receiver has an incredibly bad noise figure. (I have occasionally run into hf receivers with noise figures as high as 15 dB.) The preamp will appear to "add distortion," whereas we know from measurements that it won't go into nonlinear operation until a signal of over zero dBm is present at its input. The apparent distortion in the preamp is caused by the gain of the preamp increasing signal levels further down the amplification chain *in the receiver.*

The broadband amplifier finds its principal use as an auxiliary to test equipment – as a preamp for a frequency counter, for instance, so that the counter can be used to measure the frequency of signals too small in level to operate the counter input stages.

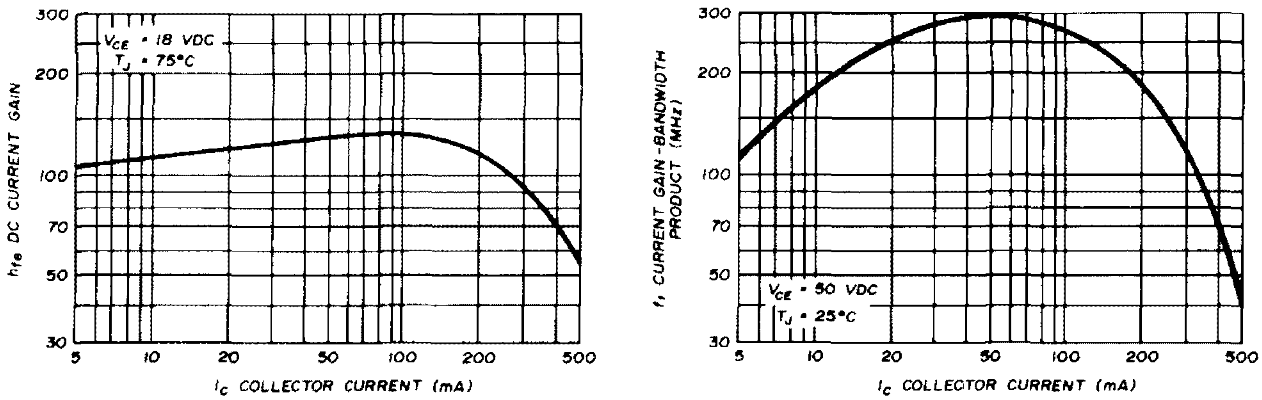


fig. 4. Dc current gain and gain-bandwidth product of the Motorola MPS-UO5 transistor used in the wide band amplifier.

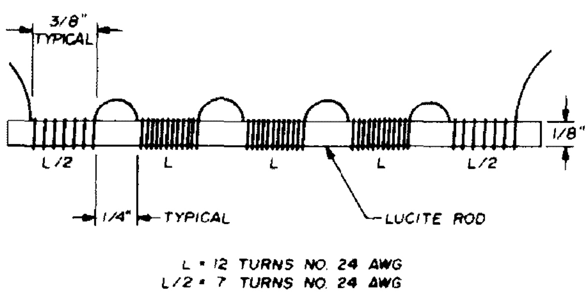


fig. 5. Construction details of the delay-line inductors.

In building up systems of available circuit blocks, the 50-ohm input and output, 18-dB gain block is quite handy. It can be used to deliver the +7 dBm local oscillator input required to drive the L port of a double-balanced hot-carrier diode mixer, for instance; or it can be used to amplify the signal going into the R port or out of the I port of such a mixer.

There are many other uses to which the broadband amplifier is uniquely suited, which involve swept frequency reception and transmission. These techniques are not in general legal for hams to use on the air, but can be useful in test and measurements on the bench.

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4. P. Beneteau and L. Blaser, "A 175 Mc Distributed Amplifier Using Silicon Mesa Transistors," Fairchild Application Note APP-14/2, 1961.

ham radio

or a scope, feed the antenna system and note the position of the meter needle or amplitude of the scope display. Note the thermo-milliammeter reading. Now insert resistance in series (always keeping the voltage constant) until the current decreases to one-half. The inserted R now equals antenna R.

In the absence of a thermo-milliammeter, use an rf voltmeter across the antenna. Again, keeping exciter voltage constant, add resistance until the rf voltage across the antenna is one-half. ($R_{ant} = R_{added}$.) We used the field-strength section of an swr meter for our rf voltmeter. We unscrewed the telescoping antenna and connected a 20k resistor in its place then connected the assembly across the antenna to ground. We adjusted for full-scale reading with no added resistance and used a vtvm across the exciter. Holding $V_{exciter}$ constant, we added resistance until the meter across the antenna-to-ground circuit read half-scale. In the case of the 40-meter antenna, we came out with a 100-ohm resistor in series with a 33-ohm resistor. The previous bridge measurement was 142 ohms. And that's about as close as you are going to get. (Actual measurement of the two resistors was 131 ohms.) A dc milliammeter, resistor, capacitor, and rf diode will work in either or both positions of the above (see **fig. 2**).

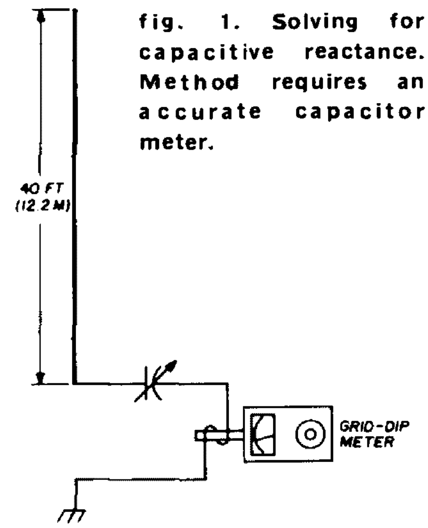

fig. 1. Solving for capacitive reactance. Method requires an accurate capacitor meter.

The accuracy of this kind of measurement hinges on the accuracy of the instrument used for making the "half" measurement and the added series resistors. Ours measured 131 ohms on dc. Carbon resistors might be slightly higher on 7 MHz. Accuracy of the exciter voltmeter is not important, as it merely has to be kept constant. Fig. 3 shows the setup.

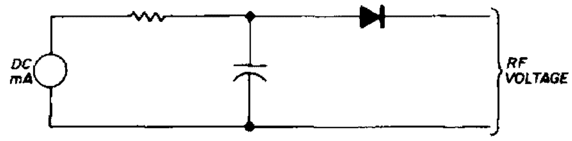

fig. 2. Simple circuit for measuring rf voltage.

This measurement method will give you a ballpark estimate near enough so that applying the results to the impedance-matching data in reference 1 will give unity swr with a bit of tweaking.

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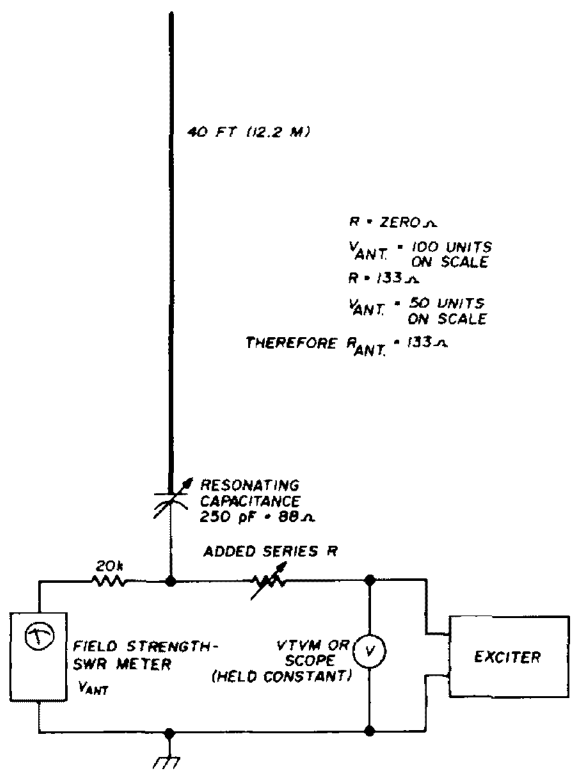

fig. 3. Test equipment for determining antenna resistance (40-meter example).

vertical antenna radiation patterns

The effect of height on vertical antenna radiation patterns

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

Various antenna articles have shown that it is possible to use almost any height vertical, from very short ones to ones which are relatively tall. For our purposes this range of heights includes electrical lengths (h/λ) from an h/λ of about 0.1 to 0.3. These ratios correspond to antenna heights from about 30 to 80 feet (9.1 to 24.4 meters) at 3.8 MHz. In this article I will compare the vertical radiation pattern of verticals in this height range. For purposes of this discussion I will assume that the *horizontal* radiation pattern is omni-directional and of no further interest.

Some antenna articles have indicated that the low-angle vertical radiation pattern for tall vertical antennas is much better than for short ones. This is a situation that I'll examine in this article, and will show that on the basis of comparison usually used, the low-angle vertical radiation patterns for both short and tall vertical antennas are essentially identical. For antennas taller than 0.3λ there is some improvement at certain radiation angles but also degradation in field strength at other radiation angles. Data on this will also be provided.

The vertical radiation angles of interest will be in the range from just above the horizontal (zero degree) through angles of up to 30 degrees. These are the radiation angles of interest for DX work on the 80-meter band. Although the far distant pattern will have a drop in signal strength near zero degree because of ground losses, primarily beyond the antenna radial system, the vertical radiation pattern of signals launched *near* the antenna and radial system is the topic of discussion in this article.

To compare the vertical radiation patterns of verticals of various heights, it is necessary to calculate the signal strength for each antenna at a fixed distance from each antenna at several radiation angles between zero and 30 degrees. It is also necessary at the same time to take into account various other factors, some of which are the same for each antenna, and some which vary with antenna height.

Equation 11-95 from page 314 of John Kraus's book, *Antennas*[1] includes all of the factors necessary to carry out the task just outlined, and his equation applies to the situations being considered here. The equation, with slight changes in notation for clarity, is

$$E(a, r) = \frac{60}{r}\sqrt{\frac{W}{R_{loop} + R_{ground}}}$$

$$\frac{\cos(\beta h \sin a) - \cos \beta h}{\cos a} \text{ volts/meter}$$

This equation gives the vertical radiation pattern for a vertical of height h, at an elevation (radiation) angle a, at a distance r meters, for a power W watts, for resistances R_{loop} and R_{ground} ohms. This situation and some of these terms are illustrated in fig. 1. Not shown is the antenna current distribution, a current element, or other items used to derive the equation for E, the field strength in volts per meter.

Now consider the various terms in the equation. The number 60 is a constant, so will be the same for all of the calculations. Since I want to calculate the field strength, E, at some arbitrarily fixed distance, I'll let r = 1 mile, which is (5280/3.28) = 1610 meters, so this term is also fixed at this value. Antenna books often use a distance of 1 mile and a power output of 1000 watts as a basis for comparison of antennas.

The term W is the power in watts going into the antenna radiation resistance and into the ground resistance. I'll use W = 600 watts to be representative of the amateur situation with a power input of 1000 watts. W is the transmitter power output less any losses in your antenna tuner, in the feedline between tuner and antenna matching network, or in the antenna matching network. For this article, I will neglect network losses and ground losses. Their influence on field strength and the radiation pattern will be considered in a later article.

There are two resistance terms involved, R_{loop} and R_{ground}. R_{loop} can be obtained from my previous article[2] by means of the equation

$$R_{loop} = R_{base} \times \sin^2 \beta h$$

The resistances R_{base} and R_{loop} are small values for short antennas, are both equal to 36.56 ohms for $h/\lambda = 0.25$, and are larger values for $h/\lambda = 0.3$ or greater. R_{ground} is the earth ground loss resistance. The ground losses are more important for short antennas where R_{loop} is small.

For example, if $R_{loop} = R_{ground}$, then half of the power, W, will go to the antenna to be radiated as a useful signal, and half will be used to heat up the ground which, of course, is not our objective and field strength will be affected accordingly. The last term in the equation is

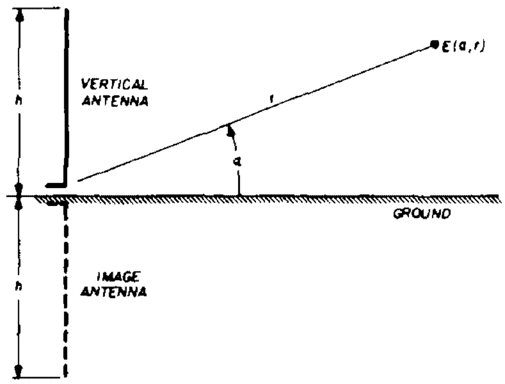

fig. 1. Vertical antenna radiation. For a discussion of the terms and their use in the field-strength formula, see the text.

$$\frac{\cos(\beta h \sin a) - \cos \beta h}{\cos a}$$

This term depends on $\beta h = (2\pi/\lambda)h$, and a, the vertical radiation angle. For a given frequency, λ is constant, so βh depends directly on the antenna height, h.

This term is plotted in some antenna articles and is then used as the basis to claim that tall verticals are much better than short verticals, since this portion of the equation does, by itself, favor tall verticals as far as low-angle radiation is concerned. It is, however, not realistic nor correct to consider only this term and to neglect the other terms discussed earlier. Now consider why using only the last term is incorrect.

In more simplified but correct terms the Kraus equation is

$$E \propto \sqrt{\frac{W}{R}} f(h, a)$$

The $\propto$ right after the E says that E is

proportional to both $\sqrt{W/R}$ and to f(h, a). Here R = R_{loop}

$$\frac{W}{R}=\sqrt{\frac{I^2R}{R}} = I \text{ or } E \propto I\, f\,(h, a)$$

This form for E, being proportional to the current, I, and f(h,a), is often given in antenna books. If only f(h,a) is plotted, it assumes a constant current, I. But since power, W, is constant, and the resistance, R, is smaller for short antennas, a constant current is not consistent with the equation $W = I^2R$ for our situation. Perhaps a better way of saying this is that to use $E \propto f\,(h,a)$ only assumes a constant current, rather than a constant power. For the amateur situation, a constant power is a much more realistic assumption.

For a constant power, $\sqrt{W/R}$ goes up as h/λ becomes smaller since R becomes smaller for short antennas. At the same time f (h,a) becomes smaller, and the two terms tend to offset each other, which is why E does not change much when constant power is assumed.

A simple non-mathematical argument about why E is essentially the same in both cases would be worthwhile. For a short antenna, E at a distance is due to the contribution of a few current elements, each with high current, while the signal from a tall antenna is produced by many current elements, each with lower current.

E (a, r) was calculated in a few minutes time on an HP-35 pocket electronic calculator for h/λ of 0.1, 0.2 and 0.3, and for a = 0°, 10°, 20° and 30°. The results are given in **fig. 2.** This graph shows that for constant power the height of radiator makes very little difference in the vertical radiation pattern. This is the essence of the entire article.

The results here are in agreement with those in most antenna books and articles, such as in Kraus, *Antennas,* page 317, figure 11-36,[1] or in Harmon, *Proceedings of the IRE,* January, 1936, pages 42-44.[3] Both of these show the sky-wave signal patterns from verticals of various heights for a fixed distance and power as being similar to **fig. 2** presented here.

table 1. Field strength for various height vertical antennas in millivolts per meter at one mile, power output = 1000 watts (multiply these values by 0.78 for 600-watts power output).

height		vertical radiation angle					
degrees	wavelength	0°	10°	20°	30°	40°	50°
360°	1.0	0	0	160	230	225	160
270°	0.75	0	0	0	140	210	225
230°	0.64	276	230	150	0	60	0
190°	0.528	246	230	200	130	70	50
180°	0.50	236	235	200	130	70	50
90°	0.25	195	190	180	160	140	110
very small		186	190	180	160	140	110

Important assumptions used here were that W = 600 watts was a constant, and the R_{ground} = 0 ohms. In a later article I'll develop R_{ground} for various radial systems, and will also show the effect of network losses reducing input power to the antenna. The reduced power and non-zero values of R_{ground} will reduce the field strength of a short vertical as compared to a tall one.

What is important is that field strength is not better for a tall vertical due to the f(h,a) term alone, but that the entire Kraus equation and correct assumptions

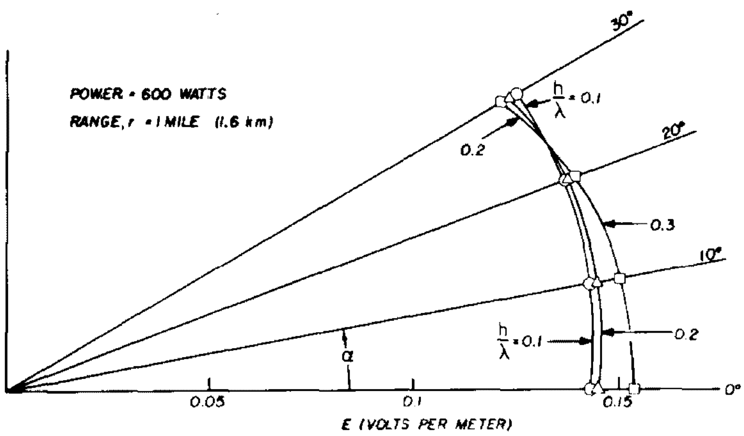

fig. 2. Vertical radiation pattern vs the height of a vertical antenna. Height in this graph is given as a ratio of the operating wavelength, h/λ

and data must be used to make a valid comparison.

This still doesn't answer the basic question of what height antenna to use or which height is better. Since the pattern versus height doesn't really matter, the important factors turn out to be earth and network losses and bandwidth. Once we know about earth losses it will be possible to answer the original set of questions and select an antenna height.

However, from a radiation standpoint, I have shown that a 0.1λ vertical (or less) is essentially the same as one 0.3λ or $\lambda/4$ wavelength tall. Furthermore, a $\lambda/2$ vertical is inferior to a $\lambda/4$ at radiation angles above 30°. It is 0.8-dB better at 10° and 0.4-dB better at 20°.

A $5/8\lambda$ vertical is inferior to a $\lambda/4$ at radiation angles above 20°. It is 0.7-dB better at 10°.

A 1λ vertical has very little radiation below 30°, but is 1.2-dB better than a $\lambda/4$ at 30°. Thus, you would have to put up a 250-foot (76.2-meter) vertical at 3.8 MHz to get about 1-dB gain for medium distances, and with poor DX performance.

There are cases where tall verticals are useful. For example, Ballantine showed back in 1924[5] that a vertical antenna approximately $5/8\lambda$ tall provided the maximum ground-wave signal, which is important for broadcast stations (most broadcast stations use a vertical somewhat less than $5/8\lambda$ tall to reduce nighttime sky-wave interference problems).

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4. Edward C. Jordan, Keith G. Balmain, *Electromagnetic Waves and Radiating Systems*, 2nd edition, Prentice-Hall, New York 1968.
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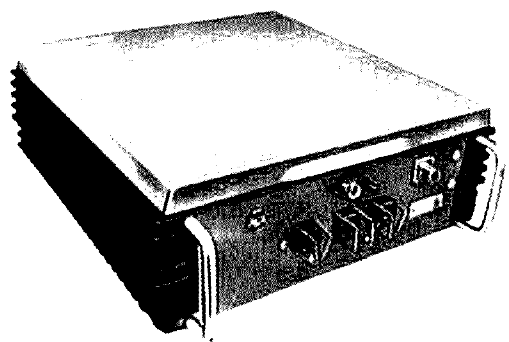


CW regenerator

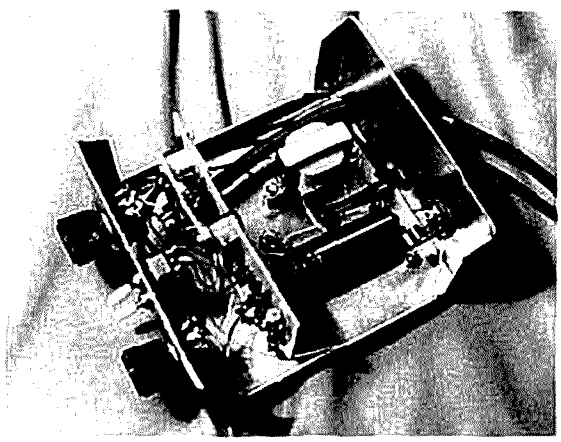
for interference-free communications

Here's a slick CW regenerator circuit that uses a phase-locked loop to provide a virtually QRM-free, single-frequency output

C.R. Lewart, R.S. Libenschek, WB2EAX, 54 Fredric Drive, Ocean, New Jersey 07712

The circuit described here should be of interest to all CW enthusiasts. It is a new approach to the old dilemma – how do you listen comfortably to a CW station deeply imbedded in noise, hash and interference? The way this is achieved is to detect one particular CW transmission and key an independent oscillator with the received signal. The input and output frequencies may be different. The result is similar to a very narrow CW filter. The main difference between the regenerator circuit and a passive CW filter is the flexibility of the active circuit and the virtually single-frequency output. Such a clean, interference-free output cannot be achieved with a passive filter.

The circuit presents an improvement over a passive filter in a certain signal-to-



Chassis layout of the CW regenerator. Vector circuit board is mounted vertically, near the front panel. Power supply components are located at the rear.

noise range. Because the active circuit requires a minimum threshold signal voltage to operate, it will fail for extremely poor signal-to-noise ratios. Relative to other means its performance is best for intermediate to good signals. The circuit was tested by the authors with both Allied-Radio Shack SX-190 and Heathkit SB100 receivers. Both receivers provided sufficiently stable and strong CW signals at the earphone or speaker jacks to operate the CW regenerator.

The CW regenerator circuit consists of a signal amplifier, a narrowband frequency detector and trigger, an oscillator, a gate and an output amplifier (fig. 1). The details of the circuit are shown in fig. 2. The incoming signal is amplified by the transistor Q1 and its output coupled to the phase-locked loop U1. This integrated circuit has two states. In the absence of a triggering signal, its output (pin 8) presents a high impedance to ground. In the presence of a triggering frequency, f_O, the output presents a low impedance to ground.

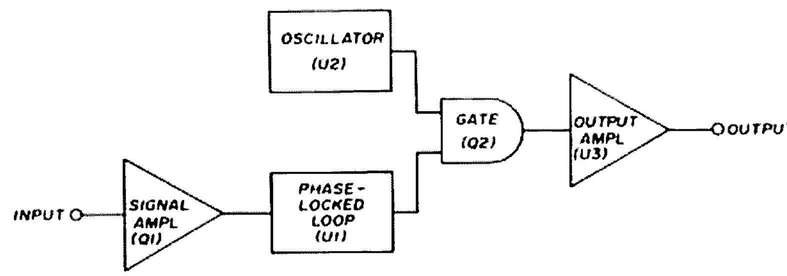


fig. 1. Block diagram of the CW regenerator. A complete schematic is shown in fig. 2.

$$f_o = \frac{1.1}{(R_4 + R_5)C_3}$$

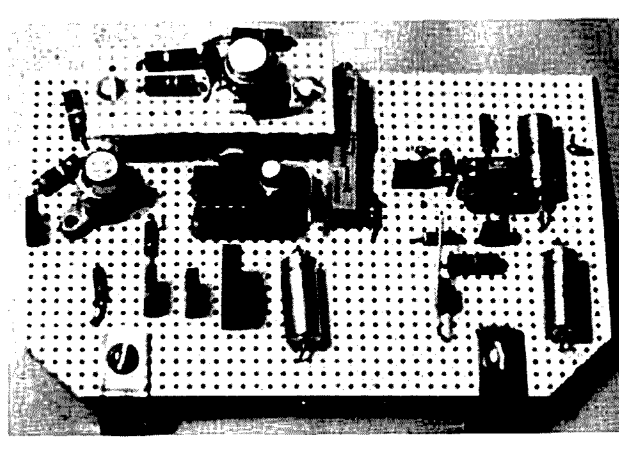
The bandwidth of this circuit is determined by capacitor C4 and is approximately 10% of the center frequency with selected values.[1] Integrated circuit U2, another phase-locked loop IC, is used as an independent oscillator; its frequency is determined by R9 and C8. The oscillator output is gated by the output of U1. The gate itself is a p-channel mosfet, Q2. Potentiometer R15 adjusts the dc voltage at the output of the gate to the same value as the dc voltage at the input. This minimizes transients. The output amplifier, U3, is a 1/4-watt audio amplifier IC used to drive earphones or speaker. Resistors R6 and R20 and capacitors C6 and C13 are used to decouple the individual stages of the circuit.

circuit development

The circuit has evolved through a number of stages. Our original idea was to use U1 only and to either trigger its internal oscillator or let it operate a low power buzzer. The results were only partially satisfactory. The dc transients caused considerable breakup of the output signal. The next step was to add a relay and later an fet gate to combat the dc transient problem. This resulted in some improvement but we found that the internal oscillator would cause transients while locking on the input signal. The use of a separate oscillator helped this situation. As the next step, input and output amplifiers were added to provide simpler interfaces.

construction

Most of the components, with the exception of the power supply, are contained on a 2½x5-inch Vector board. Although we could have used a printed-circuit board, it was found that point-to-



Vector circuit board for the CW regenerator. U2 is located on small subassembly. Q1 is center left with U1 and Q2 to the right.

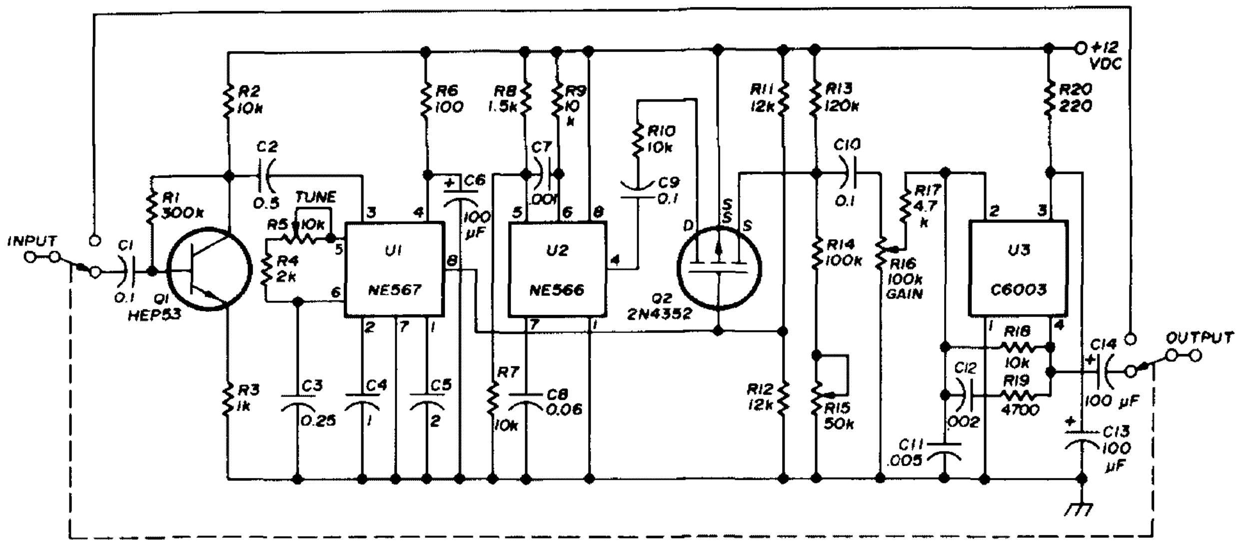


fig. 2. Schematic diagram of the CW regenerator. Integrated circuit U1 is a Signetics NE567 phase-locked loop, U2 is a Signetics NE566 function generator and U3 is a Motorola C6003 audio power amplifier.

point wiring on the Vector board afforded the easiest type of construction. All the transistors and ICs are socket mounted, with U1 and Q2 sharing a 16-pin DIP socket. The trimpot shown in the photograph is R15. This control, although not requiring frequent adjustment, might be better placed on the rear panel of the cabinet. Parts placement on the circuit board is not critical.

The homebuilt cabinet is a simple 5x5½x3-inch aluminum cabinet of U-type construction. Many types of commercially-available cabinets are suitable. The Vector board is mounted vertically toward the front with the power supply components at the rear. The control at the left on the front panel is the *tune* control, R5, while R16, the *gain* control, is at the right. The toggle switch at the bottom center is the power on-off switch. The bypass switch is located at upper center.

adjustment

Adjustment of the CW regenerator is exceedingly simple as there is only one control to adjust, the gate balance control, R15. With no signal input and a high-impedance voltmeter connected between the gate and source of the mosfet, Q2, adjust R15 for zero voltage. This is the only non-operating adjustment required.

In operation, center the desired CW signal in the passband of the receiver with the function switch in the *bypass* position and the receiver audio gain control in the center position. Place the main switch in the *regenerate* position and adjust the *tune* control, R5, to lock in the signal. This should be an absolutely pure regenerated signal with the quality of a code oscillator. There will be no QRM or noise. With two or three signals in a pileup, the *tune* control will allow you to completely peel off the unwanted QRM. The *gain* control should be adjusted to provide a comfortable signal level.

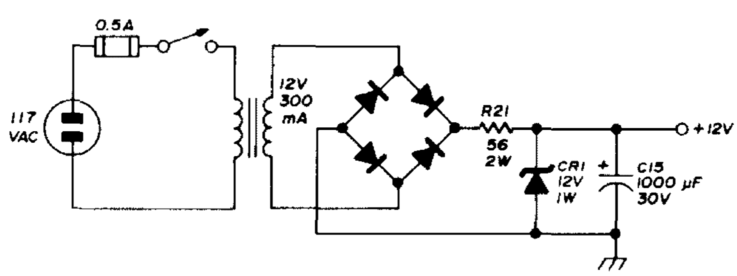


fig. 3. Power supply for the CW regenerator.

reference

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ham radio

the ham notebook

printed circuit standards

Recently there has been a lot of interesting home-built equipment described in the amateur radio magazines. Frequently a firm offers ready-made printed-circuit boards to aid those readers who wish to duplicate one of these designs. Also, a lot of ready-made and kit-form equipment is now being offered in printed-circuit form, which the buyer can enclose as he chooses. Although these printed circuits come in all different sizes and shapes, the designer probably originally planned to mount the board on spacers inside a Minibox of convenient size. After you have built a few projects of this type you find yourself with a small pyramid of various sizes of Miniboxes, each with a power cord and other dangling wires. The organization and esthetics of the hamshack would be much improved, not to mention the ease of constructing new projects, if a few standards could be adopted for printed-circuit construction.

Interestingly enough, there already exists a *de facto* standard in this area: a card width of 4.5 inches (11.4 cm). There are probably a dozen manufacturers offering card cages which fit into a 5.25x19-inch (13.3x48.3-cm) rack panel space and accept cards up to 4.5-inches (11.4-cm) wide. A number of these card cages are adjustable for smaller widths but in any case this width is by far the most popular. For breadboarding purposes companies like Vector and Vero offer a wide variety of boards with various hole and etch patterns, with or without edge-connector lands, and all in the 4.5-inch (11.4-cm)width. Hence the first request is that printed-circuit designers try to confine themselves to a 4.5-inch width and that manufacturers of boards center narrower patterns on a 4.5-inch space. If the user wants to mount a smaller pattern in the smallest possible space he can always trim off the excess width.

Most amateur equipment designs require rf shielding so open-card-cage-style construction is not suitable. For a single project the Minibox enclosure is entirely satisfactory for most of us. With a 4.5-inch (11.4-cm) pattern the designer might want to use a 5.5-inch (14-crn) board width, which allows a half-inch (13 mm) on each side of the pattern for the screws and spacers used to hold the board to the cover of the box. For larger installations, both of the manufacturers previously mentioned offer a nice system of cases which mount in something similar to a card cage. An example of this kind of construction is illustrated in WA6JYJ's RTTY speed-converter article in the December, 1971, issue of *ham radio.*

These cases come in several different standard heights, one of which is designed for 4.5-inch (11.4-cm) cards. Of course, if you want to take advantage of this system of mounting multiple units you must standardize on one height. This packaging system is by no means as inexpensive as the Minibox, and it isn't as easily available from the corner radio store, but the cost is still small compared to the cost of what goes into the boxes and the items are readily available from

the factory by mail order. If cost is really an obstacle you can home-make something that looks just as nice and has the same standard dimensions but costs a lot less.

Having suggested that a standard 4.5-inch (11.4-cm) width will help us to build equipment more easily and uniformly, how about the other printed-circuit board dimension? There is really no need to standardize in this direction, for nothing says that cards mounted in an open cage or in cases have to all be the same length. Again, looking at the manufacturers' catalogs we see that several different lengths are offered, but nothing much over nine inches (22.9 cm). Hence, if you can hold the length of a board to less than this, and choose cases or card cages big enough to hold a board at least this long, you can accommodate projects of all different sizes.

A card cage might look funny with different-length cards sticking out of it, but you can cover them up with an attractive, hinged, front door. Then the only problem is how to extract a short card from between two longer ones; that is easily solved with a couple of holes in the front edge and a U-shaped piece of coathanger wire with the tips bent to form hooks to engage holes in the cards.

Finally, what about external connections to the cards or cases? The most popular scheme for cards in an open cage is the familiar card-edge connector that mates with lands etched on the board. These are available in an enormous variety of sizes and features but it is surprising how often you see connectors with 22 contact positions on 0.156-inch (4-mm) centers. These are available with 22 contacts on one side or double-sided with 44.

There are many options open for the external connections to cards mounted in boxes or cases. You could just run cords out the back of the case, or have a terminal strip on the back. This doesn't ease the old tangle of wires, but at least it doesn't show from the front. A more attractive possibility is to have a connector on the back of the case which mates with a connector attached to the frame holding the cases. Then an encased unit can be removed for inspection with no dangling wires; and it can be operated in the removed position by means of an extension cord running from the plug on the case to the socket in the frame. Or, the socket can be left unattached to the frame so that the case can be pulled out for servicing without the use of an extender. Again, there are many suitable kinds of connectors that are inexpensive, such as the ordinary octal plug. My own preference is for the 22-position card edge connector. This allows encased cards to be easily intermixed in the same frame with open cards. A card in a case has the usual edge-connector lands which project through a slot in the back of the case to engage the connector. This makes it unnecessary to run a lot of wires from some other kind of connector to the card inside the case.

There are lots of ready-made experimental cards designed for this kind of construction that can be used either in cases or in open guides. Unfortunately, these are pretty high priced as compared to plain Vectorboard but they might occasionally be worth their price in the time they save. It's certainly a convenience to be able to pull out a piece of equipment and plug another in its place for testing with no wires or cables to bother with.

In summary, here are three recommendations for uniform packaging:

1. Design printed circuit boards for a standard width of 4.5 inches (11.4 cm).

2. Keep the other dimension of the board to about 9 inches (22.9 cm) or less; if a project is too large for this size, use more than one board.

3. If a printed-circuit edge connector is appropriate, use the kind having 22 contact positions on 0.156-inch (4-mm) spacing.

Jim Haynes, W6JVE

new products

solid-state ssb transceiver

The new Atlas-180 ssb transceiver recently introduced by Atlas Radio offers a number of unusual features in an amateur high-frequency ssb transceiver. Most obvious, of course, is its extremely small size: 9½ inches (24.1 cm) wide, 3½ inches (8.9 cm) high, and 9¼ inches (23.5 cm) deep. Weight is a mere 7 pounds (3.2 kg). Packed into this small package is a complete all solid-state 180-watt PEP ssb and CW transceiver for the 20-, 40-, 80- and 160-meter bands (crystal oscillator accessory available for MARS). The Atlas-180 is ideal for mobile or portable operation since it operates directly from a 12- to 14-Vdc source, negative ground, drawing 200 to 400 mA in the receive mode, 16 amps peak on transmit. An optional ac supply is available for home station use.

The Atlas-180 uses modular construction, including plug-in circuit boards, for ease of service and maintenance. Connectors on the rear of the transceiver are designed to plug into the mobile mounting bracket, or into the AR-117 desk-top ac power supply, making transfer or removal a simple operation.

The receiver in the new Atlas-180 features sensitivity of less than 0.5 microvolt (typically 0.25 μV) for 10-dB signal-plus-noise to noise ratio while providing excellent immunity to overload and cross modulation. The input signals are converted directly to a 5520-kHz i-f without any preamplification. Selectivity is provided by a 5520-kHz crystal lattice filter with a 6-dB bandwidth of 2.7 kHz and a shape factor of 1.7. Ultimate filter rejection is 110 dB. Receiver image rejection is greater than 60 dB. Included in the receiver is an internal speaker, S-meter and 100-kHz crystal calibrator.

Receiver and transmitter frequency control is provided by a highly stable vfo circuit. The tuning dial is calibrated in 5-kHz increments and is easily interpolated to 1 kHz. Tuning rate is 15 kHz per revolution.

The transmitter has several interesting features including a broadband design which eliminates transmitter tuning and single conversion from the i-f to the output frequency. Included in the circuit is ALC and infinite vswr protection. Input power is 180-watts PEP on ssb; output is 80 watts PEP minimum (100 watts PEP typical). Unwanted sideband suppression is better than 60 dB at 1000 Hz, and carrier suppression is more than 50 dB below peak power. Intermodulation distortion is approximately 30 dB down and image outputs are more than 40 dB below rated peak output. Harmonic output is more than 35 dB below rated peak power.

Accessories available for the Atlas-180 include the AR-117 table-top 117-volt ac power supply and mobile mounting bracket (deluxe plug-in and non-plug-in models available). Automatic VOX (CW semi-break-in), phone patch and other accessories will be announced in the near future. For more information, write to Atlas Radio Inc., Post Office Box A, Carlsbad, California 92008, or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

antenna tower protection system

A new system introduced by the Towtec Corporation automatically guards against tower and antenna damage from high winds. Called *Towrgard*, the system continuously monitors your local wind velocity. When the wind velocity reaches or exceeds the changeable, preset 35 mph limit (even on gusts) the system will automatically lower your motorized tower. Towtec also manufactures electric hoists for manual crank-up towers.

The *Towrgard* system includes a computer/controller, a wind sensor and a bottom limit switch for automatic motor turn-off when the tower is nested. For more information on this *automatic* system, write to the Towtec Corporation, 118 Rosedale Road, Yonkers, New York 10710, or use *check-off* on page 94.

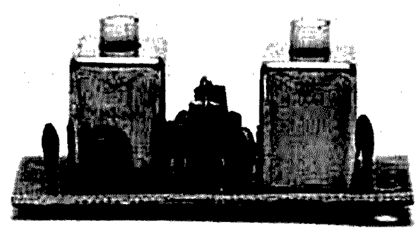
cambion catalog

The new Cambion XQ Components Catalog illustrating various Cambion (CTC) components now available through retail outlets is now being offered at no charge. Many popular items are illustrated, such as terminals, jacks, plugs, handles, battery holders, IC sockets, IC breadboards, coils and rf chokes.

Cambridge Thermionic Corporation, better known as CTC, has been a manufacturer and supplier to industry of precision electronic components for over thirty years and has just recently made its Cambion products available to the amateur and experimenter. Industrial, commercial, military and aerospace engineers use more than 20,000 different components in the Cambion line. Through this catalog the amateur and experimenter now have available the same quality and reliability.

For further information on the new Cambion XQ Components Catalog, write to Cambion, Department XQ, 145 Concord Avenue, Cambridge, Massachusetts 02138, or use *check-off* on page 94.

vhf preamp



Hamtronics, Inc. has announced several new products for the vhf buff. Of primary interest is a new improved version of the well-known cascode 6- and 2-meter preamps described in articles in March, 1972, and January, 1973, issues of *ham radio*. The new units are 3/4-inch wide x 2-inches long x 1-inch high (19x51x25-mm) so they will fit into almost any receiver or transceiver. New fets used in the preamp provide more gain and better noise figure while retaining the inherent stability of the cascode circuit which has made the Hamtronics preamp so popular. Amateurs like this unit because they can tune it easily without worry about neutralization.

With several thousand preamps now in the field, hams are reporting that even newer transceivers are improved by adding a low-noise preamp ahead of the front end. Units are still $6.00 in kit form or $10.00 wired and tested, postpaid in the USA. Club prices are still available. Models are also available now for any frequency from 20 to 240 MHz, including 10-meter Oscar reception.

Other new products to be announced soon include a lower-cost, more-compact scanner device for up to four channels; a receiver kit for a-m reception on aircraft frequencies or 6 or 2 meters for net use; and 450-MHz receivers, preamps, and transmitters for fm. Also, kits are now available for the popular 5-watt audio amplifier, using an integrated circuit, as written up in an article appearing in September, 1972, *ham radio*.

For more information, send a self-addressed, stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612, attention Jerry Vogt, WA2GCF, or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

digital voltmeter

Now a low-priced, laboratory-quality digital voltmeter is available from MITS, Inc. The model DVM 1600 measures alternating and direct current in five ranges from 0.1 mA to 1 amp. Ac and dc voltage is measured in four ranges from one volt to 1000 volts. Measurement of resistance is in six ranges from 100 ohms to 10 megohms.

The resolution in low ranges for voltage is 10 mV; for current, 10 mA; and for resistance, 1 ohm. Dc voltage accuracy is ±.5%. All other measurements are accurate to ±1%. Input impedance for dc voltage measurements is 10 megohms; for ac voltage, 1 megohm.

The DVM 1600 also features auto polarity which automatically displays polarity and magnitude without probe reversal. Other features include a regulated power supply and 100% overrange capability on all ranges. Power requirements are 115/230 Vac, 50/60 Hz, 20 watts.

The MITS DVM 1600 digital voltohmmeter is available in an easy to assemble kit form ($89.95) or factory assembled ($129.95). Warranty on the assembled model is one year on parts and labor. Kit warranty is ninety days on parts. For more information, write to MITS, Inc., 6328 Linn Avenue, NE, Albuquerque, New Mexico 87108, or use *check-off* on page 94.

short circuit

transistor curve tracer

There are several errors in the schematic for the transistor curve tracer published in the July, 1973, issue, page 53. There should be a 0.02-μF capacitor connected from the collector of Q1 to the collector of Q2; the value of C4 should be 0.2 μF, *not* 0.02. Also, CR9 should be a 1N457. When building this unit be sure that the circuit ground (the common ground for Q1, Q2, Q3, Q5 and Q7) is separate from the chassis grounds used on the oscilloscope jacks (J1 and J2).

85 cents

focus
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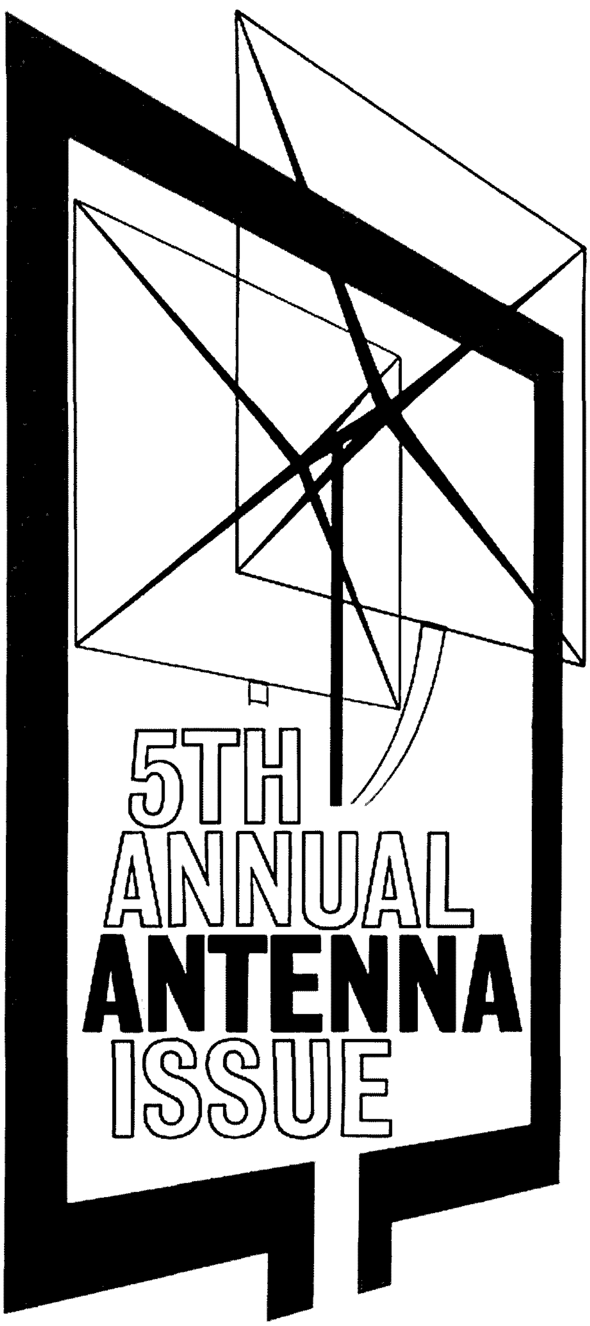
ham radio magazine

MAY 1974



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Hilda M. Wetherbee
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advertising manager

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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contents

a second look

by jim fisk

Ever since the military services first started using wireless back around the turn of the century their engineers and radio officers have been trying to find a way to use trees and other natural objects as antennas. Many methods were tried, including the simple expedient of pounding nails into trees and connecting the transmission line to them. Nothing worked. It was found, in fact, that trees actually absorbed most radio waves and that thick forests made radio communications all but impossible over any but short ranges.

It wasn't until relatively recently, in 1969, that any real progress was made. At that time an intensive research program was begun at the Army Electronics Command's research laboratory. In the four years since the program began a team of scientists under the direction of Dr. Kurt Ikrath have developed a system for making effective use of trees and other natural objects as antennas. The key, of course, is the coupling of rf energy into and out of the object. The Army Command accomplished this through a flexible, toroid-shaped hybrid electromagnetic antenna coupler called a *Hemac* which is formed in a circle around the tree as shown in the photograph.

This tree is radiating a signal on 10.8 MHz. The transmitter is coupled to the tree through a matching network (in the box on the tree) and the toroidal Hemac. (U.S. Army photo.)

The Hemac operates as the primary of a leaky rf transformer; the tree or other core object acts as a single-turn secondary. A variable tuning and impedance-matching circuit is used with the Hemac to provide a good match to 50-ohm transmission lines. Dr. Ikrath reports that the Hemac-tree antenna requires less skill to tune than the short whip on the Army's PRC-74 radio set. Nor is the Hemac's use limited to trees – it has been used successfully with such man-made objects as light standards, window frames and helicopters. In one case two Hemac coils were used with trees, four meters apart, as a high-frequency phased array. By changing the phase difference between the 4.65-MHz signals driving the two trees, engineers were able to vary the radiation pattern.

As might be expected, frequency is very important to the operation of the system. In a forest environment the dense underbrush scatters or absorbs short wavelengths. Long wavelengths seem to work best as absorption is much less. The optimum operating frequency also depends upon the object used as the antenna. A 100-foot tree, for example, works best in the 80-meter range. Who will be the first to put this idea to work on Field Day?

Jim Fisk, W1DTY
editor-in-chief

fixed log-periodic beam for 15 and 20 meters

A simple, low-cost wire beam that provides lots of gain per dollar

G.E. Smith, W4AEO, 1816 Brevard Place, Camden, South Carolina 29020

Since retiring from business in 1970 I have been designing, building and testing fixed, high-frequency, log-periodic beam antennas. I had long wondered why amateurs had made so little use of these very excellent beams which are used extensively by commercial, military and government communicators.

To date, over fifteen log periodics have been erected and thoroughly tested here. Most of these have been horizontal doublet log periodics for the 20-, 15- and 10-meter bands, as shown in plan form in fig. 1. Some have been for 40, 20 and 15. Two vertical monopole log periodics (with ground plane) were tested on 40 and 80, giving 10-dB gain with the low angle of radiation best suited for DX.

Since most of these log periodics cover a 2:1 or one-octave bandwidth, e.g., 7-14 or 14-28 MHz, they are rather long for the average amateur antenna farm. Two of the log periodics tested here for 14-30 MHz were one with a 40-foot (12.2-meter) boom length, having 8-dB gain, and one with a 70-foot (21.35 meter boom, giving 10-dB gain as comparec with a doublet at the same height. M\ longest log periodic is 100-feet (30.5 meters) long, has 17 elements, and give: 12- to 13-dB gain. The swr for these antennas is relatively low over the 20- 15- and 10-meter bands, generally no exceeding 2:1 with 1.1:1 to 1.4:1 typical

Since these log periodics are wire beams, they are quite inexpensive con sidering their gain. They are also quite easy to build. The material cost usually runs from $15 to $25 each, less feedline and masts or other supports.

As a result of on-the-air discussions while testing and comparing these anten nas, and from several previous articles or log periodics, I receive many requests for information on the smallest possible log periodic to cover at least 14 through 21.5 MHz. The inquirers generally want gain equal to or better than the average ham beam.

two-band log periodic

A log periodic having 8-dB gain and meeting these requirements, erected in a space 40 by 40 feet (12.2 x 12.2 meters), will be described here. This antenna, which is beamed south, has been in use for three years. It gives excellent performance. Also included is a list of materials, approximate total price and brief assembly information.

Fig. 2 shows the layout for this simple 7-element log periodic for 20 and 15 meters. Its bandwidth or frequency coverage is 14 to 21.5 MHz. Theoretically it provides 8-dB gain in the forward direction. However, when compared with a dipole at the same height, reports in the direction of its beam generally show a 10-dB increase over the doublet. When an

S-9 report is received while using the doublet, a "20 over 9" is often received after switching to the log periodic. An equivalent S-meter increase at this end usually confirms the reports received.

At any rate, the reports received during three years of use give this log periodic a conservative 10-dB gain in its forward direction. It is mounted approximately 40 feet (12.2 meters) above ground. The measured swr over the two bands is:

14.0 MHz	1.1:1	21.0 MHz	1.01:1
14.1 MHz	1.1:1	21.1 MHz	1.01:1
14.2 MHz	1.02:1	21.2 MHz	1.05:1
14.3 MHz	1.02:1	21.3 MHz	1.15:1
14.35 MHz	1.01:1	21.4 MHz	1.25:1
		21.45 MHz	1.3:1

If even greater gain is desired and the space is available, a 9-element log periodic having a boom length of 57.3 feet (17.48 meters) can be used (**fig. 3**). This is actually the 20-15-10 log periodic of **fig. 1** with several of the short (10-meter) elements deleted, which reduces its bandwidth to 14 to 21.5 MHz for operation on 20 and 15 meters. This log periodic should give 10- to 11-dB gain on 20 and 15 if its height is at least a half wavelength above ground for 20, about 33 feet (10 meters).

If space is available for an even longer antenna, **fig. 4** illustrates another log periodic which should give 11- to 12-dB gain. The latter two antennas (**fig. 3** and 4) have not actually been tested here, but the gain figures were taken from the 3-band equivalents for 20-15 and 10. Deleting the 10-meter section should have little effect on performance for 20 and 15.

The front-to-back ratio of these log periodics will be approximately 10 to 12 dB, not as good as a Yagi, but with its other advantages over a Yagi the log periodic is well worth consideration.

Fig. 5 shows the **fig. 2** log periodic as it would look suspended from four 40-foot (12.2 meter) masts. These can be inexpensive telescoping TV masts, available at any TV shop, or 45-foot (13.7 meter) wooden poles. Used poles are available from power companies in some areas at a very reasonable price. The TV masts will, of course, require guying, but wooden poles can usually be unguyed for the smaller log periodics. The log periodic of **fig. 2** as used here has its rear end supported by the roof and the short forward end by two cedar trees. This tree-roof configuration just happened to be perfect for suspending this log periodic 40 feet (12.2 meters) above ground, beamed due south.

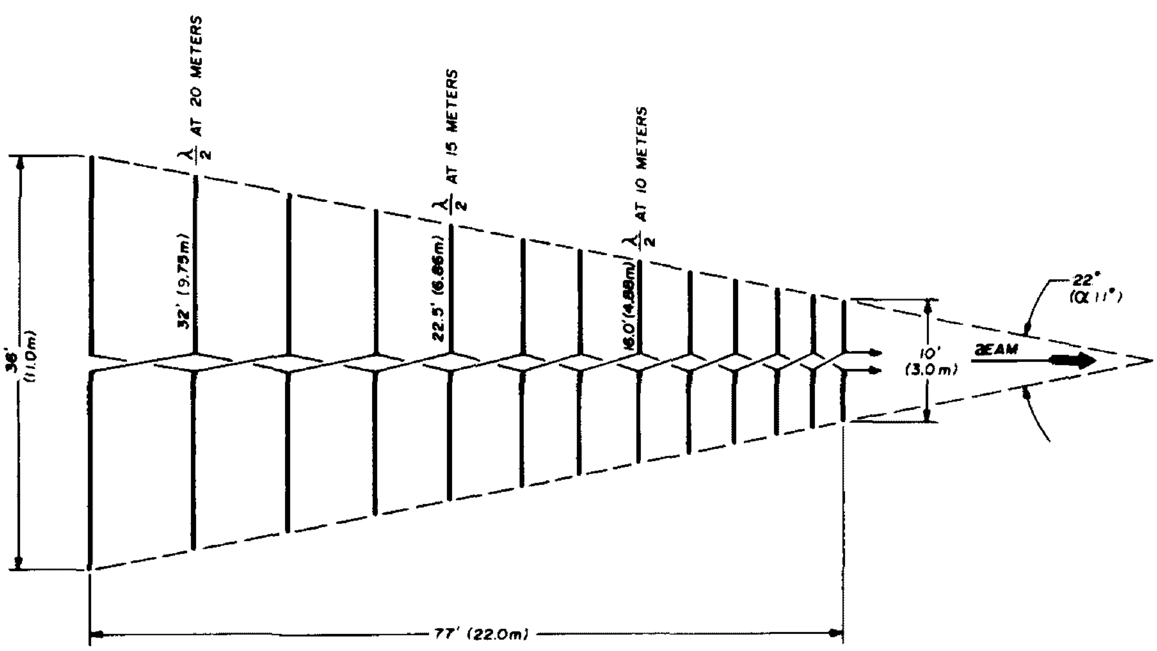

fig. 1. Plan view of a typical log periodic beam for 14-30 MHz. With 13 elements and a boom length of 72 feet (22 meters) this antenna provides 10-dB gain over a reference half-wave dipole for 10, 15 and 20 meters. Note the feedpoint transposition of alternate elements — if this is not done, the antenna will not work.

If any of these two-band log periodics can be suspended at least a full wavelength above ground (approximately 66 feet on 20 meters) they will no doubt provide a lower angle of radiation. This is better for DX and, in effect, will show greater gain, especially on 20. The highest I have been able to use here has been 60 feet (18.3 meters), and considerable improvement was noted on that particular antenna compared to the others installed approximately 40 feet (12.2 meters) above ground.

The beam width of a log periodic is generally about 90 degrees, which is good for a fixed beam required to cover a particular continent or a certain part of the U.S. From this location, the antenna beamed west covers the entire west coast and also Australia.

construction

Fig. 6 illustrates the physical layout of the 7-element antenna showing location of the home-made Lucite insulators and the suspension of the seven elements between the two nylon sidelines or catenaries. Note the transposition of the feedline to alternate elements, a must for a log periodic.

Fig. 7 is a drawing of the insulators, which are made from 1/4-inch (6-mm) Lucite. The end insulators have only the two end holes, while four holes are drilled in the center insulators. The two end holes are for fastening the center of the elements, while the two toward the center, spaced 1½ inches (3.8 cm), are for the two-wire center feeder. The center insulators serve two functions: First, they separate and space the two-wire open feeder, and second, being secured to the feeder at the points specified in fig. 2, they space the elements at the required distances so the antenna will function as a log periodic.

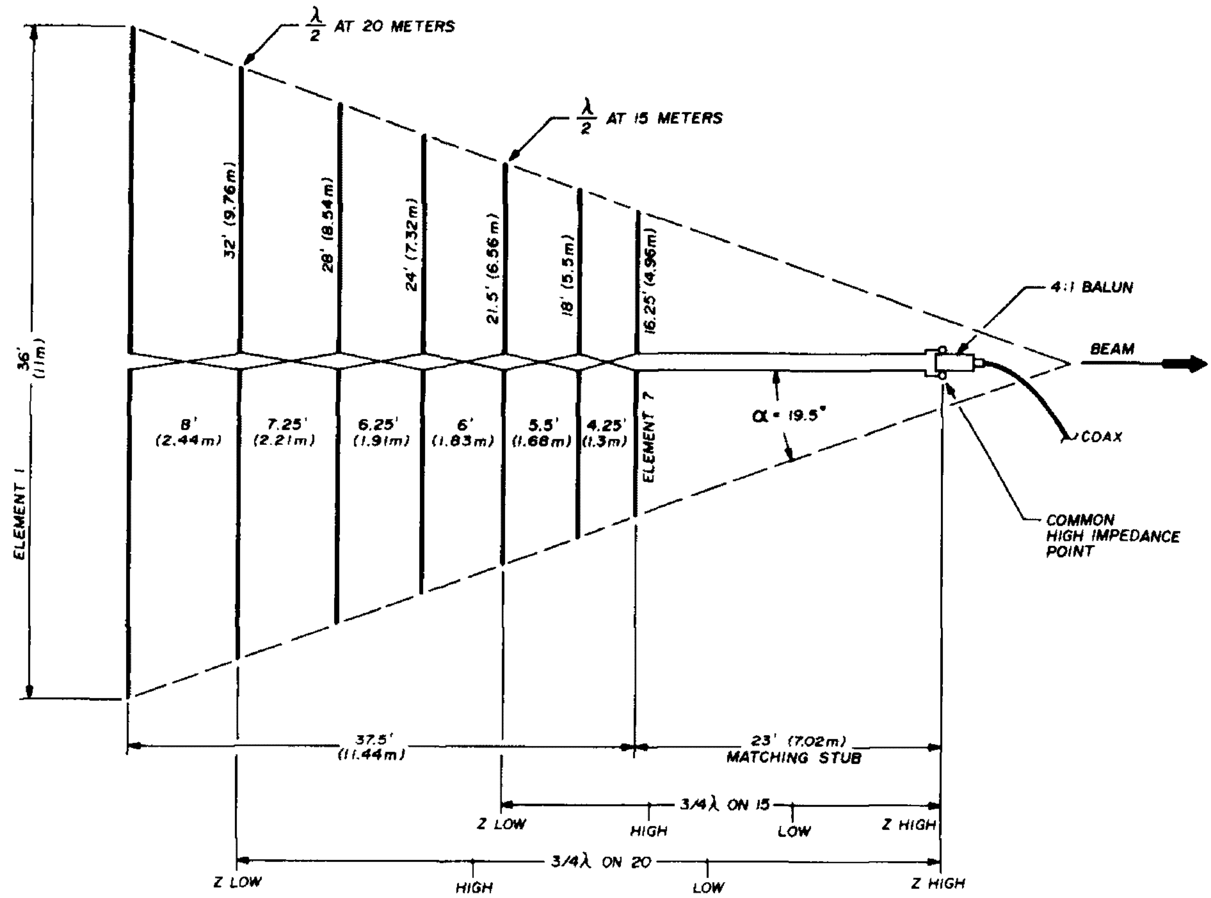

fig. 2. Dimensional drawing of a practical log periodic for 15- and 20-meter operation that requires only a 40- by 40-foot (12- by 12-meter) space for installation. Though its theoretical gain is 8 dB, the author's version of this antenna has consistently shown 10 dB or better gain over a reference dipole at the same height. Note feedpoint transposition of alternate elements.

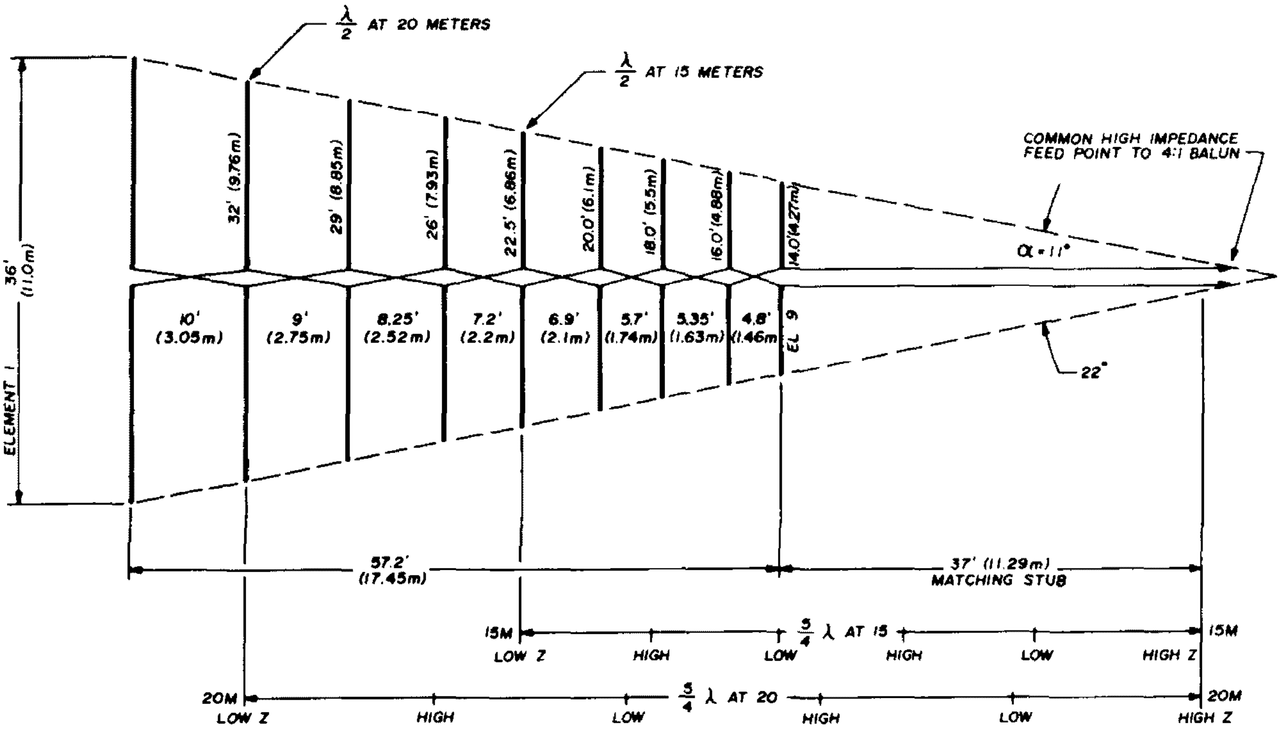

fig. 3. Dimensional drawing of the log periodic of fig. 1 with the 10 meter elements omitted. Though its 57.2-foot (17.45-meter) boom length may be prohibitive in some installations, those who have the room for it should find a 2- to 3-dB improvement compared to the smaller antenna of fig. 2.

Note that six egg-type ceramic insulators are used in place of Lucite for the long rear element, 1, and for the short forward element, 7. This is recommended due to the additional strain on these two elements, which must support the remaining five elements and the weight of the center feeder.

Note that the two-wire center feeder is extended 23 feet (7.0 meters) beyond the short forward element, 7. This provides a common impedance match on both 20 and 15 meters. A 4:1 balun is used to provide a match between the coax transmission line and the antenna. This 23-foot stub would not be necessary if a

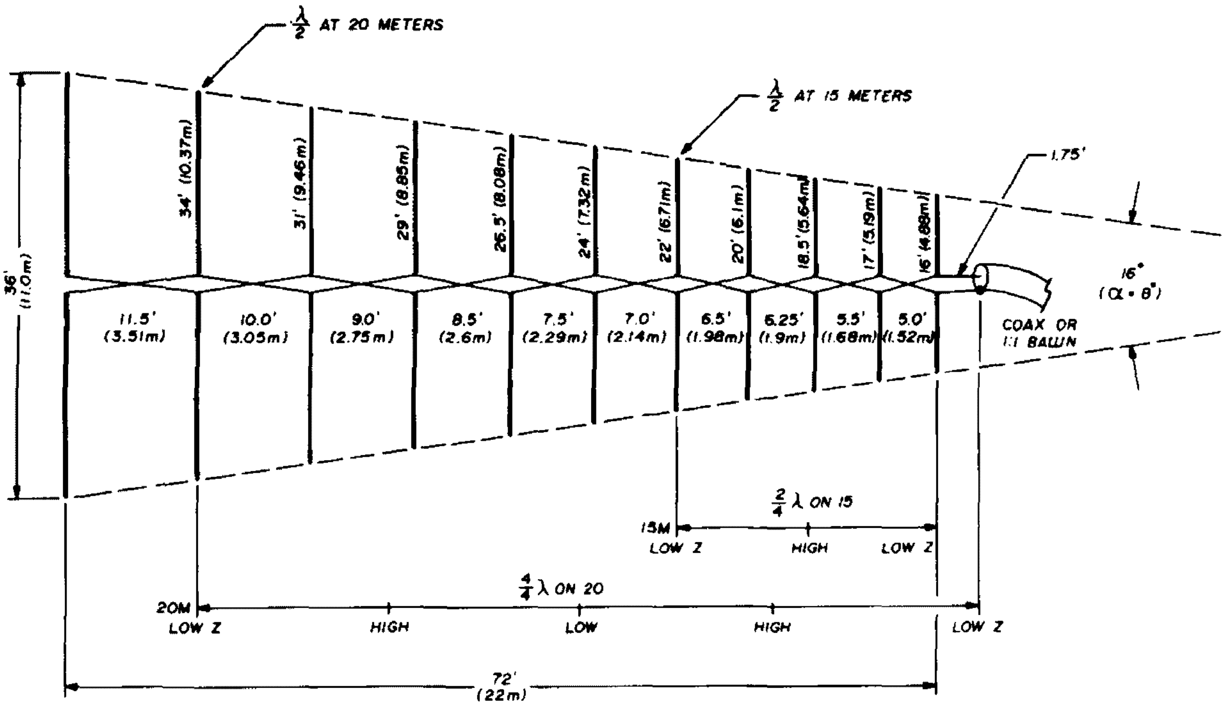

fig. 4. Dimensional drawing of an even longer two-band log periodic. This 11-element array should provide at least 12-dB gain over a reference dipole.

tuned line was used between the shack and the log periodic. However, this would require a tuner or *Match Box* at the equipment end. The use of the stub plus the 4:1 balun eliminates the need for the antenna tuner.

It may be of interest that the above feed method has a very desirable feature. Element 2, which is the driven or active radiator element on 20 meters, is spaced approximately 3/4 wavelength from the common feedpoint. Element 5, the driven or active element on 15 meters, is also

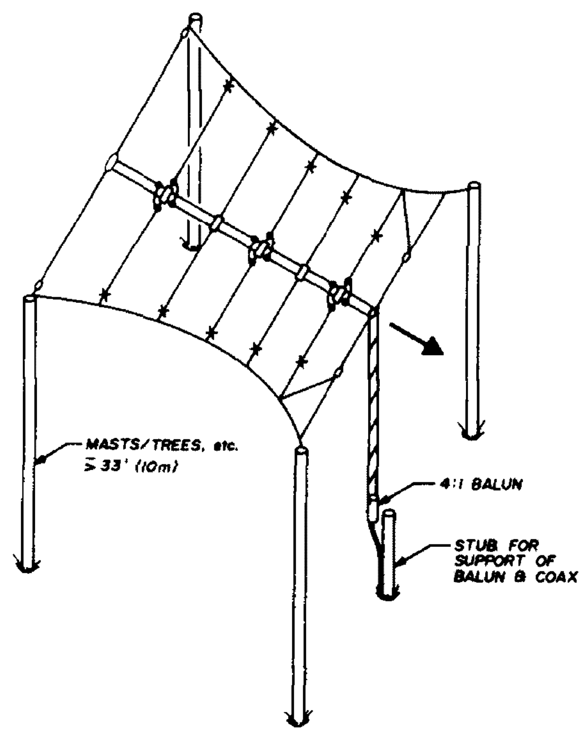

fig. 5. TV masting or used utility poles provide an easy means for putting up a wire log periodic antenna such as the seven-element array of fig. 2.

3/4 wavelength from the balun. Since the active element for a discrete frequency within the bandwidth of the antenna is 3/4 wavelength from the common feedpoint, the center feeder plus the 23-foot (7.0-meter) stub act as an impedance matching line and the impedance at the common feetpoint is relatively constant on either 20 or 15 meters.

The stub can hang down from the antenna as shown in fig. 5. A short stub mast can be used to support the balun and the coax, relieving the strain on the short-element end. Several additional

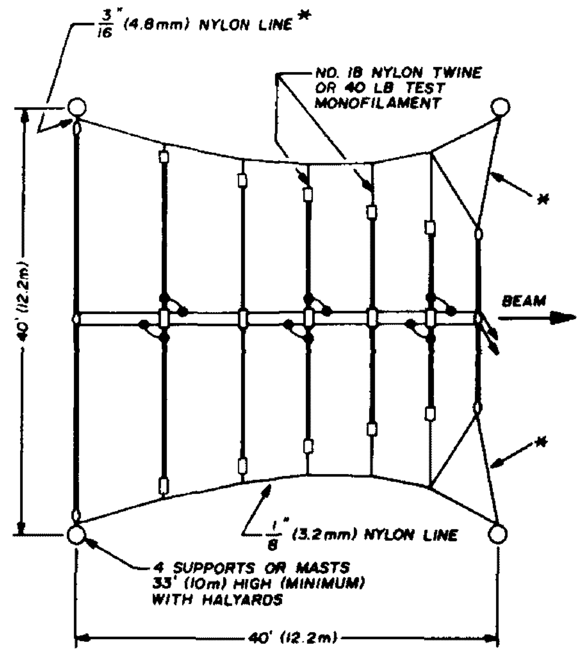

fig. 6. Mechanical layout showing suspension of log periodic of fig. 2. Egg insulators are used on the first and last elements, while the Lucite insulators detailed in fig. 7 are used elsewhere.

Lucite spacers should be used along the stub to keep the two wires separated and parallel. Spacers about every five feet (1.5 meters) are suggested.

log-periodic theory

The log-periodic beam can be considered as a multi-element, broadband, uni-

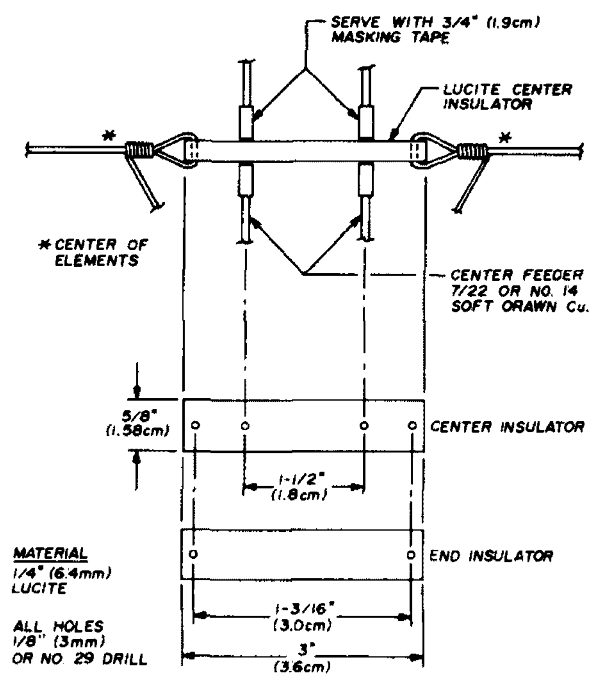

fig. 7. Dimensional drawings of Lucite insulators used at centers and ends of center elements. Dimensions and material are not critical.

table 1. List of materials needed for building the log-periodic antennas described here (not including masts or coax).

item	7 elements (figs. 6 & 2)	9 elements (fig. 3)	11 elements (fig. 4)
Antenna wire (elements)[2]	190 feet (58 meters)	225 feet (69 meters)	290 feet (89 meters)
Antenna wire (center feeder)[1,2]	125 feet (38 meters)	196 feet (60 meters)	160 feet (49 meters)
Ceramic egg insulators	6 each	6 each	6 each
End insulators (Lucite)	10 each	14 each	18 each
Center insulators (Lucite)	5 each	7 each	9 each
Nylon line, 1/8 inch (3 mm)	100 feet (30.5 meters)	170 feet (52 meters)	235 feet (72 meters)
Nylon line, 3/16 inch (5 mm)	25 feet (7.6 meters)	25 feet (7.6 meters)	25 feet (7.6 meters)
Nylon twine, no. 18[3]	1 roll	1 roll	1 roll
Total cost	$17.00	$22.00	$26.00

notes

1. The center feeder should be 7/24 copper (not copper-clad — too stiff) or other stranded wire so the two wires will remain parallel and not kink.
2. The author used number-15 aluminum electric fence wire on many of his log periodics to reduce weight and cost. Caution must be used when working aluminum, and care must be taken to avoid contact between dissimilar metals which can cause electrolysis. Aluminum is not recommended for use near sea coast or in a polluted environment.
3. 40 lb. test monofilament fish line can also be used, eliminating need for end insulators.

directional, end-fire array. The design formulas have appeared before in an amateur publication,[4] which although it covered only vhf log periodics, also applies to high-frequency log periodics. The design formulas are quite complex, so unless a computer is available it is suggested that the dimensions given in this article be followed.

Table 1 is a list of materials for the short 20-15 meter log periodic, **fig. 2**, as well as the two longer models of **fig. 3** and **4**.

summary

Anyone who has the space to put up one of these excellent log periodic antennas for 20 and 15 will be pleased with the results. Considering its moderate cost, its gain in the forward direction is about equivalent to adding a big linear. It is equally helpful in reception in the desired direction, giving the same gain in receiving. It also has a very pronounced diversity effect, especially important during bad signal fading. Of all the various beam antennas tried here during the past three years, the two-band log periodic for 20 and 15 has been the simplest and easiest to construct. I believe it gives more "miles per dollar" than almost any other amateur beam.

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ham radio

parabolic reflector antennas

Complete construction details for a large parabolic reflector for uhf — Australian style

T.R. Naughton, VK3ATN

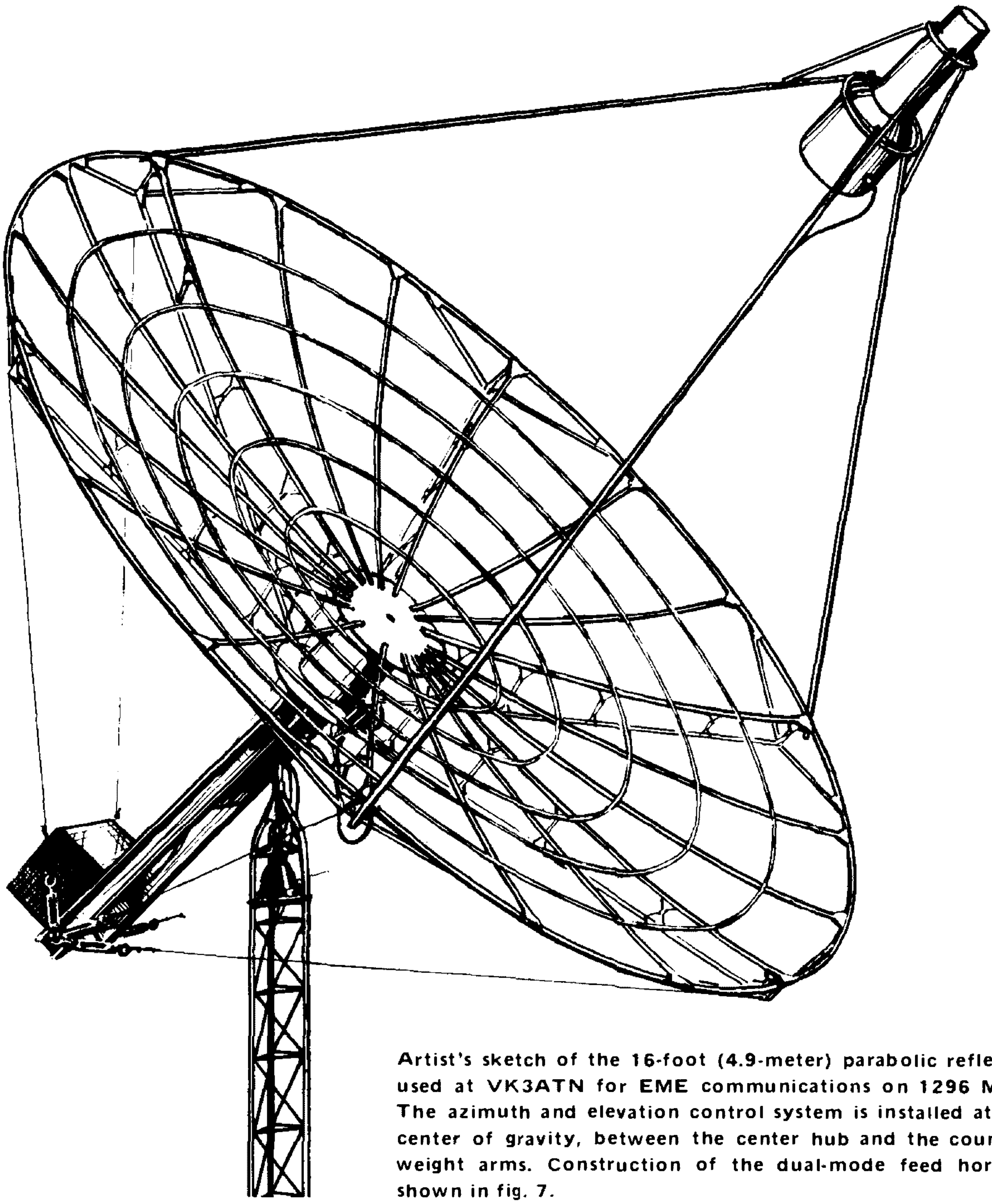
There should be little doubt in the minds of avid uhf experimenters that the antenna is a most important part of his station. This is especially so for long-path DX and EME amateur communications where the most important characteristic of the antenna is its ability to efficiently direct all of the available transmitter power in a preferred direction — its effective gain. Since antennas are passive reciprocal devices they perform the same unique high-gain function in the receiving mode as well. There should also be little doubt that, in the uhf range and higher frequencies, the parabolic reflector is the most efficient, high-gain antenna. Those experts in antenna technology may debate this last statement; however, for home construction without critical measurement and adjustment, the parabolic reflector antenna has no equal.

This article describes fabrication techniques using readily available materials which can be used in the construction of relatively large parabolic reflectors with exceptional rigidity, strength and moderate weight. Construction is mainly of round aluminum tubing with gusset plates and pop rivet fasteners. The conductive reflector surface is aluminum screen or galvanized iron mesh, also readily available.

Although much detail will be described, this article is not intended to give a *cook-book* design. Using the same general construction techniques described here the size of the reflector can range from 10- to 30-feet (3 to 9 meters) in diameter. The design is not unlike many commercial reflectors. However, unlike commercial construction, no aluminum welding is required and weight is kept to a minimum.

feed systems

The efficiency of a parabolic dish antenna is highly dependent on the feed antenna and on the surface accuracy of the reflector. A very efficient feed is the dual-mode, small-aperture antenna whose radiation characteristics determine a focal-length-to-aperture-diameter ratio of 0.6 for the reflector.[1] However, this particular feed antenna, because of its physical size, is not practical for frequencies below about 1000 MHz. A suitable feed for 432 MHz is the NBS gain



Artist's sketch of the 16-foot (4.9-meter) parabolic reflector used at VK3ATN for EME communications on 1296 MHz. The azimuth and elevation control system is installed at the center of gravity, between the center hub and the counterweight arms. Construction of the dual-mode feed horn is shown in fig. 7.

standard whose radiation characteristics are very similar to the dual-mode aperture antenna. Both feed systems are described at the end of this article. Although these feeds are shown as linearly polarized, they may be adapted for circular polarization. This is recommended to eliminate Faraday fading for moonbounce communications.[2]

surface accuracy

Surface accuracy of a reflector antenna affects the phase of the aperture wave. Variations from an exact paraboloid surface cause scattering of energy in directions away from the main beam and, thus, lower the gain. The reflector design presented here, together with the dual-mode feed system, will produce an effective gain of

$$G_{eff} = 0.65 \left(\frac{\pi D}{\lambda}\right)^2$$

where D is the reflector physical diameter, λ is the operating wavelength in the same length units as D, and the surface

accuracy, everywhere on the surface, is within 0.025λ of a true paraboloid.

construction details

Construction of the reflector consists of four parts: one, a central hub from which construction begins and which is also used to mount the antenna; two, an array of identically trussed ribs which are

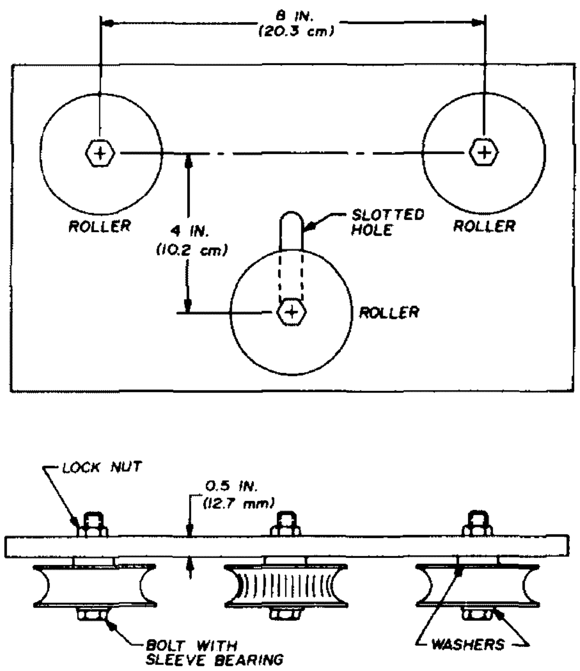


fig. 1. Rolling mill construction. Curved grooves in the rollers should be slightly larger in radius than the tubing to be bent. Middle roller should be knurled or roughened if this roller is driven. Slot for the middle roller mounting bolt is to permit changing bending radius.

attached to the hub and form the parabolic shape; three, a series of circumferential rings which join the ribs and provide support for the conductive surface material, and four, the conductive surface. Mounting of the feed, although important to the use of the reflector, will be considered separately from construction of the paraboloid.

The aluminum tubing used in this construction is a half-hard or 6100 alloy which can be easily bent. Bending can be done several ways. Two methods suggested here are by a rolling mill, which must be constructed, or by a special tool called a *hickey* which is used by electricians for bending conduit. These may be bought from most electrical supply houses for about $5.

The rolling mill produces a smooth arc bend and is very desirable for bending the circular rings which support the reflector surface material. The rolling mill used in the construction described here is shown in **fig. 1**. With this mill it is important that the groove in each roller is a bit larger than the tubing which is to be bent. Also, it is necessary to have a secure and resettable position for the middle roller to set the bend radius. A hand crank or other drive mechanism attached to the middle roller will ease the required rolling effort. Even though it will take some time to construct the rolling mill, its use will greatly shorten total construction time and produce very uniform and repeatable bends. Since the extreme ends of tubing bent in the rolling mill will not conform to the curvature of the central portion, a few inches should be cut off at each end. Some care should be taken in rolling as the tubing may have a tendency to twist.

The *hickey* is used to kink the tubing in small incremental sections producing an arc composed of short straight sections between kinks. This method of bending is entirely acceptable for the rib and hub construction. Note that the actual reflector surface material is not supported by the ribs but by the rings which are spaced about one foot apart along the ribs.

Virtually all the fastening of tubing is done with aluminum or steel pop rivets. These are readily available in hardware stores along with a manually operated tool which properly sets the rivets. This method of fastening is quick and provides a long-lasting permanent fastener. A unique feature of pop rivets is that they may be used in blind holes as shown by **fig. 2**. Either 1/16- or 5/32-inch (1.6- or 4-mm) diameter rivets may be used for most fastening, but larger rivets may be used where heavy pieces or thick pieces are to be joined.

The center hub which I used consists of two circular rolled rings of 7/8-inch (22.2-mm) tubing about 24 inches (61 cm) in diameter and spaced 18 inches

(45.7 cm) apart. A number of cross braces are used to hold the rings together rigidly. While this construction method used only tubing and rivets, an improved hub construction would be a spool or reel design made of 1/8-inch (3.2-mm) aluminum plate and a 16-inch (40.6-cm) section of large diameter aluminum tubing. This hub design is shown in **fig.** 3.

The end discs of the hub are secured to the large tubing by means of many small aluminum angle brackets with rivets. If a large diameter tube is not available a suitable substitute would be a hexagonal section composed of flat 1/8-inch (3.2-mm) plates. Additional stiffening of this hexagonal tube at its center is required and may consist of a single band of aluminum bent to conform to the outside of the hexagon and fastened near the plate seams with rivets.

A very important part of the hub is the holes cut in the center of each end plate. The size of the holes is made to be a snug fit with a standard aluminum or steel tube available, 1½ to 2 inches (3.8 to 5.1 cm) in diameter. These holes and the tube form the axis of symmetry of the paraboloid and will be used to align the ribs and surface material as described later. The hub length should be about one-twelfth the reflector diameter; the end discs need not be larger than 24 inches (61 cm) for any size reflector.

Sets of two or three rivet holes can be drilled in the hub end plates at the radial locations of the ribs. Additionally, similar sets of holes can be drilled to mount the gusset plates used to attach the back member of the rib to the hub.

trussed-rib construction

To maintain uniformity of the ribs and accuracy of the parabolic shape it is essential that a jig be built to hold the rib pieces in position while fastening. For the 16-foot (4.9-meter) diameter reflector, an 8-foot (2.45-meter) piece of half-inch (13-mm) plywood may be used with wooden blocks to serve for alignment and holding. Alternatively, the jig may be made of angle iron, welded together. If the plywood jig is used, one edge of the plywood sheet is used as the axis of symmetry and the parabolic curve may be carefully laid out and drawn using the coordinates given in **table** 1. Draw in the remaining truss members as shown in **fig.** 3. The height of the rib at the hub end must be equal to the hub end plate

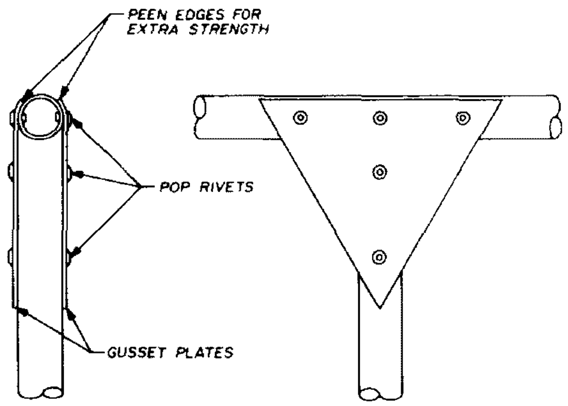


fig. 2. Gussett plate detail. Gusset plates are not necessarily all the same shape.

spacing so that the rib may be fastened to the hub with a minimum of strain.

Blocks of wood about ½ inch (13 mm) in height may now be securely attached to the plywood jig so that the truss members are held in their appropriate positions. It should be noted that the parabolic member is actually bent into a circular arc which will require a minimum of distortion to form the parabolic arc. The radius of curvature which fits this requirement is very nearly twice the focal length. Prebending of the parabolic members is done in the rolling mill or by the multiple kink method. If the latter method is employed, the kinks should occur between the locations of the circular rings. In this way the points of ring attachment will conform to the parabolic arc in the jig. These points should be marked on the rib members for ease in indexing the rings later.

The cross braces in the ribs are short pieces of the same diameter tubing as the ribs and spaced about 18 inches (45.7 cm) apart. The cross braces should be notched to fit the main members. The notching operation may be done with a

table 1. Parabolic curve coordinates for reflector diameters of 16 feet (4.9 meters), 20 feet (6.1 meters) and 24 feet (7.3 meters). Coordinates are *given for only* one-half of the parabola since the other half is identical.

reflector diameter		16 feet (4.88 meters)		20 feet (6.10 meters)		24 feet (7.30 meters)	
focal length (f)		9.6 feet (2.93 meters)		12 feet (3.66 meters)		14.4 feet (4.39 meters)	
inches	cm	inches	cm	inches	cm	inches	cm
0	0	0	0	0	0	0	0
3.94	10	0.034	0.085	0.027	0.68	0.022	0.057
7.87	20	0.13	0.34	0.11	0.27	0.09	0.23
11.81	30	0.30	0.77	0.24	0.62	0.20	0.51
15.75	40	0.54	1.37	0.43	1.09	0.36	0.91
19.69	50	0.84	2.14	0.67	1.71	0.56	1.42
23.62	60	1.21	3.08	0.97	2.46	0.81	2.05
27.55	70	1.65	4.19	1.32	3.35	1.10	2.79
31.50	80	2.16	5.47	1.72	4.37	1.44	3.65
35.43	90	2.72	6.92	2.18	5.54	1.82	4.61
39.37	100	3.36	8.54	2.69	6.84	2.24	5.70
43.31	110	4.07	10.34	3.26	8.27	2.71	6.89
47.24	120	4.84	12.30	3.88	9.84	3.23	8.20
51.18	130	5.68	14.44	4.55	11.55	3.79	9.63
55.12	140	6.59	16.75	5.27	13.40	4.40	11.16
59.06	150	7.57	19.22	6.05	15.38	5.05	12.81
62.99	160	8.61	21.87	6.89	17.50	5.74	14.58
66.93	170	9.72	24.69	7.78	19.75	6.48	16.46
70.87	180	10.90	27.68	8.72	22.15	7.27	18.45
74.80	190	12.14	30.84	9.71	24.67	8.10	20.56
78.74	200	13.45	34.18	10.76	27.34	8.97	22.78
82.68	210	14.83	37.68	11.87	30.14	9.89	25.12
86.61	220	16.28	41.35	13.02	33.08	10.85	27.57
90.55	230	17.80	45.20	14.24	36.16	11.86	30.13
94.49	240	19.38	49.20	15.50	39.37	12.92	32.81
96.00	243.8	20.00	50.80	16.00	40.64	13.33	33.86
98.43	250			16.82	42.72	14.02	35.60
102.36	260			18.19	46.21	15.16	38.50
106.30	270			19.62	49.83	16.35	41.52
110.24	280			21.10	53.59	17.58	44.66
114.17	290			22.63	57.48	18.86	47.90
118.11	300			24.22	61.52	20.18	51.26
120.0	302.8			25.00	63.50	20.83	52.91
122.05	310					21.55	54.74
125.98	320					22.96	58.33
129.92	330					24.42	62.03
133.86	340					25.92	65.84
137.80	350					27.47	69.77
141.73	360					29.06	73.82
144.00	365.8					30.00	76.20

rotary rasp of the same diameter as the tubing or with a notching tool used by welding fabricators. A wooden jig can be built to hold the tube at the appropriate angle if the rotary rasp method is used. A less pretentious but adequate notch may be vee-shaped and made by hand with a file or saw.

With all the rib members in the jig, gusset plate fastening of the butt joints is accomplished as shown by the detail drawing of **fig. 2.** The gusset plates may be predrilled and used as a guide for drilling the tubing through a single wall thickness. One hole drilled and pop riveted at a time will eliminate hole misalignment. When all gusset plates are secured on one side of the rib, the rib is

removed from the jig and turned over for fastening the second set of gusset plates. The gusset plates are cut from half-hard sheet, 22 gauge or about 1/32-inch (0.8-mm) thick. The reverse rib alignment template may be built in a similar way on the same plywood sheet.

The alignment template and support tube must be constructed accurately because they determine the ultimate accuracy of the reflector surface. This consideration is important since the reflector can be used at higher frequencies

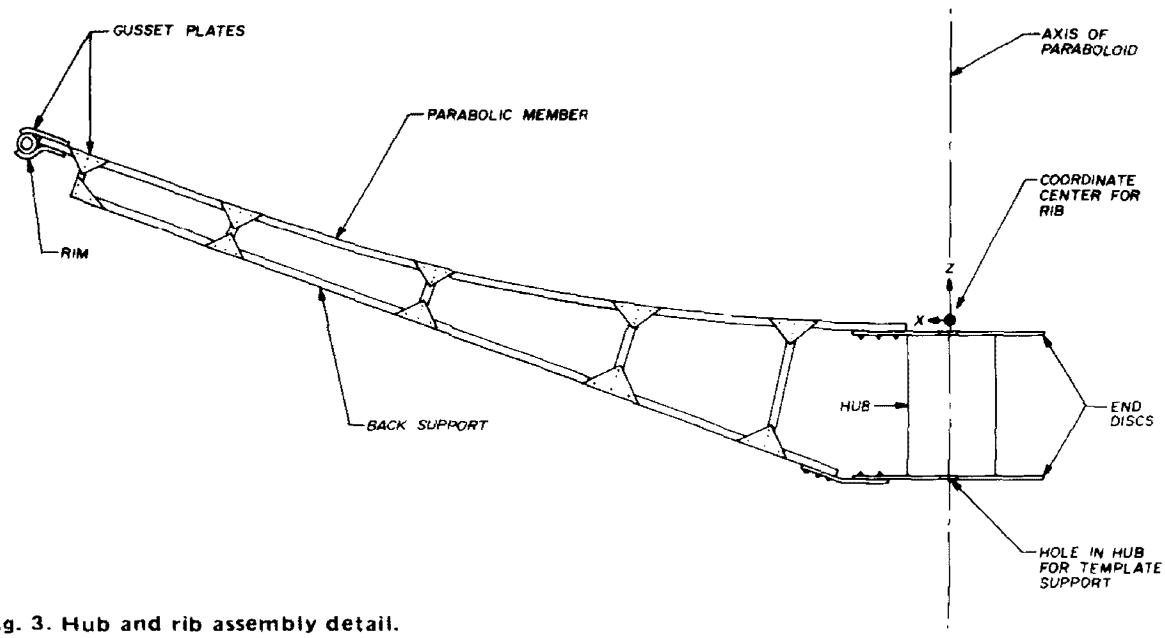


fig. 3. Hub and rib assembly detail.

if the surface accuracy is good. The alignment template support tube must fit snugly into the holes at the center of the hub.

For larger reflector diameters it may not be possible to obtain tubing of sufficient length and splices will be required. The method I used consists of taking a short length (4 to 6 inches or 10 to 15 cm) of the same tubing to be spliced (see **fig. 4**). Cut a single slot lengthwise about ¼-inch (6.5-mm) wide. Then squeeze the piece until it fits inside the splice pieces, forming a butt joint. Finally, pop rivets are used to form a permanent joint as shown by **fig. 4**.

surface support rings

Although I used 1-inch (25-mm) diameter tubing as the rim for my 16-foot (4.9-meter) reflector, it is recommended that 1½-inch (38-mm) tubing be used if possible for additional strength. The rim may be rolled to the prescribed radius bend in the rolling mill by starting with a larger radius and rolling the pieces through several times, each time decreasing the bend radius until the final value is achieved. A large circular arc may be chalk-marked on a convenient surface (garage floor, driveway, patio, etc.) for checking the sections as they are rolled. Assemble the arc sections into a circle at the final construction site using the butt splice technique except for the final splice. Leave this last splice open with some overlap so that exact adjustment of the last splice will give a good fit to the ends of the ribs.

The circular rings of 3/8-inch (9.5-mm) tubing may be rolled after the hub, ribs and rim have been assembled. These rings should be spaced concentrically with the rim starting at the rim. Approximately one-foot (30.5-cm) spacing between rings will be adequate to support the screening. At the rim a surface registration problem will occur because of the way the rim is attached to the rib ends. This problem can be resolved by either attaching a 3/8-inch

(9.5-mm) ring directly at the rim or by deliberately bending the last one foot (30.5 cm) of the parabolic member so that the rim tube will be in correct surface registration. This bend correction must be included in the rib jig before the ribs are assembled and must also include the extra kink in the parabolic member.

Since the rings will require a long section of 3/8-inch (9.5-mm) tubing, it may be necessary to splice pieces. For this small-size tubing it will be adequate to use a simple short overlapping splice riveted together. Splicing may be done after rolling, and some pre-squashing of the tubing is desirable before drilling and riveting.

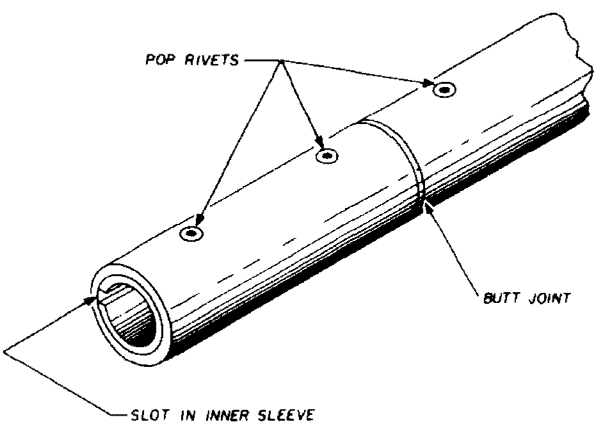


fig. 4. Method of joining sections of tubing. Use eight rivets in diametrically opposite pairs for extra strength.

assembly

Assembly starts by riveting the hub gusset plates on at the locations of the back rib members. These gusset plates should be of 1/16-inch (1.6-mm) material for extra strength. The parabolic member of the rib may now be riveted to the top of the hub and the back member temporarily clamped. Alignment of the rib is now accomplished by inserting the reverse rib template support tube into the hole at the center of the hub. The rib should be adjusted by loosening the back member clamp and allowing the parabolic member of the rib to conform exactly to the edge of the template. Now the back member may be riveted in place. Repeat this procedure for all ribs.

As the structure grows, it will be necessary to provide temporary trusses between ribs to prevent strains and lateral bending. Diametric wire or heavy cord guys between rib ends will also aid in maintaining the ribs in alingment. The reflector will be constructed with the parabolic surface initially upward.

When all the ribs are attached a survey check of rib alignment should be made with the template before the rim is attached. The rim is attached to the upper and lower surfaces of the parabolic members by means of gusset plates and rivets. Additional support for the rim may be added in the form of diagonal members from a point about 18 inches (45.7 cm) from the rib on the rim to a point near the end of the back member of each rib.

At this point in construction a number of additional single member parabolic ribs should be carefully bent, formed into parabolic shape, checked against the template or jig and then fastened to the rim and hub so that at the rim there will be a rib spacing of about 2 to 3 feet (61 to 91 cm), no more. For the 16-foot (4.9-meter) diameter dish I built, two additional single-member ribs are added between the six trussed ribs. Check these additional ribs for alignment with the template.

When forming these single-member ribs they should be marked so that the correct end is attached to the rim. The curvature is so slight that it is quite easy to reverse the rib end-for-end by mistake. The locations of the rings may now be

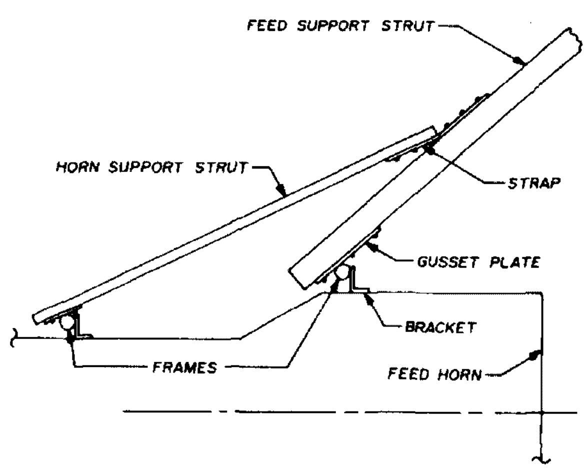


fig. 5. Partial view of the feed support detail.

marked on each rib member by first marking the locations on the template along the parabolic edge and transferring these marks to each rib.

Finally, the 3/8-inch (9.5-mm) surface-support rings are formed and riveted to the ribs. The rings must be squashed to a thickness of 5/16 inch (7.9 mm) at rivet points in order to use 5/32 x ½-inch (4 x 12.7-mm) long rivets. The rings may be temporarily clamped to the ribs to aid the assembly. After all of the rings are riveted in place a final check should be made with the template for accuracy. If errors greater than ¼-inch (6.4-mm) are found they should be corrected, even if rivets must be extracted and parts altered. If you have come this far you are to be congratulated, the rest is easy.

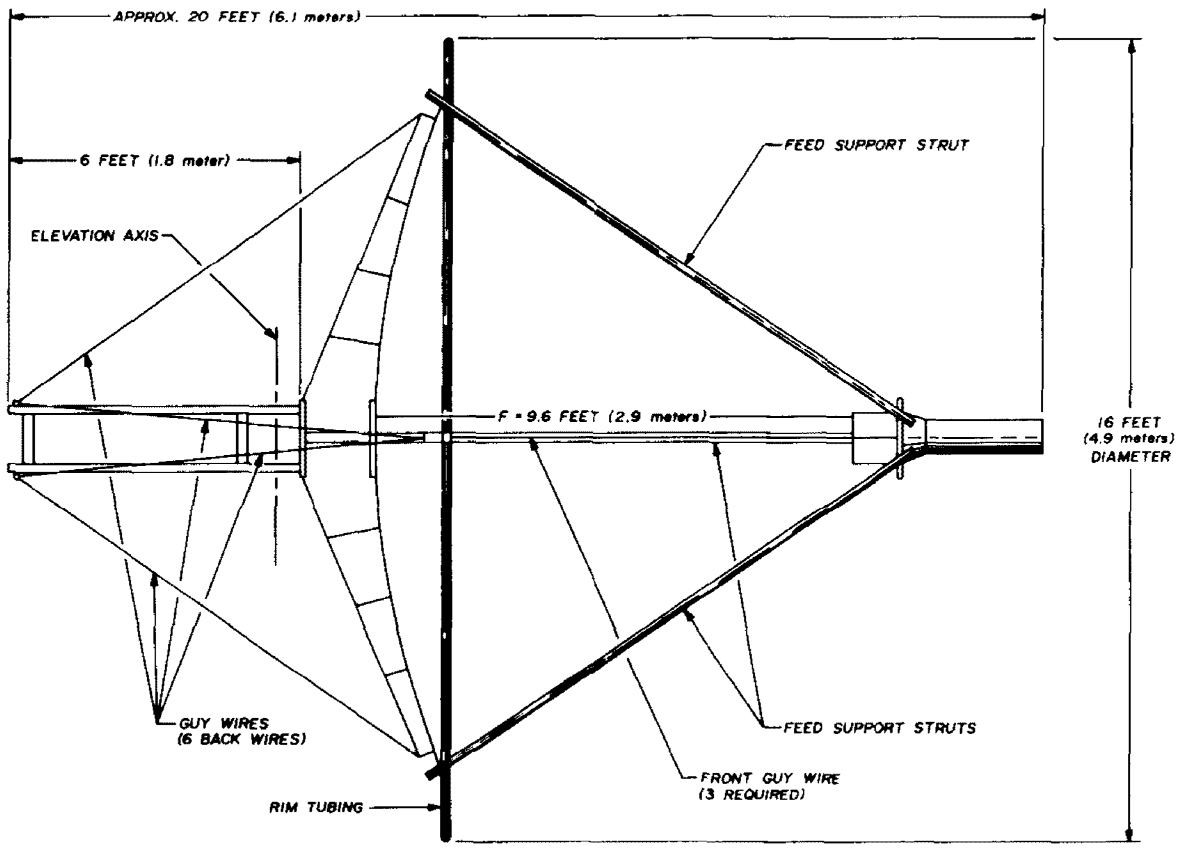


fig. 6. Skeleton plane view showing rigging and overall dimensions for the 16-foot (4.9-meter) diameter parabolic reflector.

surface construction

Surfacing consists of cutting radial length strips of screening diagonally. These triangular sections are then carefully laid over the rings between pairs of ribs in pie section fashion and fastened to the rings with wire ties. The ties should be spaced a maximum of 4 to 6 inches (10 to 15 cm) apart. Start along a ring at the middle of the pie section and work towards the rim and hub. Try to keep the screening taut with as little bowing as possible. Trim off any excess at the center to permit the template support tube to be inserted at the center of the hub. The smallest 3/8-inch (9.5-mm) support ring should be no less than about 6 inches (15 cm) in diameter. It will be more convenient to rivet the screening down where ties cannot be used. A thin washer above the screening will prevent damage if aluminum window screening is used.

At the rim, the screening should be long enough to be wrapped around the outside of the rim and secured by a spiral wrap of wire which passes through the screen near the point of tangential con-

tact. As each section of surface material is added, the edge overlap should be a minimum of 2 inches (5 cm) and held together by wire ties to prevent gaps. A last check of screening surface accuracy should be made with the template and the screening trimmed up if necessary. Pinching the mesh can be used to take up slack areas. Every effort should be made to maintain a surface accuracy of ±¼-inch (±6.4-mm) over the entire surface to assure satisfactory performance at 1296 MHz.

If galvanized iron mesh (hardware cloth) is used, the mesh size will permit inserting the tip of a pair of long-nose pliers and twisting the mesh wire to draw up specific areas where sags appear. This procedure, though tedious, can result in a very taut surface.

In the event of damage to a small surface area, an overlay of screen may be added on, provided the patch size is not less than one wavelength in its smallest dimension. Larger area damage may require replacement of an entire pie section of screening.

Another useful suggestion is to use epoxy cement to tighten up rivets which may become loosened.

feed support

The feed support is a tripod of 1½-inch (3.8-cm) diameter thin-wall aluminum tubing. For extra stiffness, and to damp wind vibrations, these tubes may be filled with foam-in-place urethane foam or billets of cylindrically-cut rigid foam. This procedure prevents diametric deformation of the thin wall and can add much to the tubing stiffness with little added weight. The foam-in-place urethane is a two-part mixture which is mixed and quickly poured down the tube with the tube in a vertical position. A wad is forced into the tubing near the middle before foaming and the foam poured in from each end.

The tripod legs are attached to the outside of the rim with short pieces of aluminum angle and bolts or, alternatively, with heavy gusset plates. The preferred locations of the legs are one near top center when the reflector is in its normal zero-degree elevation position and the other two legs spaced equally to either side of bottom center.

A frame assembly to support a 1296-MHz feed horn or higher-frequency feed systems is placed at the apex of the tripod. This is shown in part in **fig. 5.** This partial drawing shows details for one leg. The other two legs are similar but spaced around the frame in the same manner that the legs are spaced around the rim of the reflector. The ends of each leg are fastened to the frame by heavy gusset plates and rivets. These gusset plates are approximately diamond shaped. Additional strength can be provided for the thin-wall legs at their ends by a similar technique used in splicing. In this case the split sleeve inside the leg tubing serves as additional wall thickness.

The feed holding frame may be a circular ring rolled from heavy-wall, soft-alloy aluminum conduit or a square of rigid tubing assembled with corner gusset plates. To support the rather long circu-

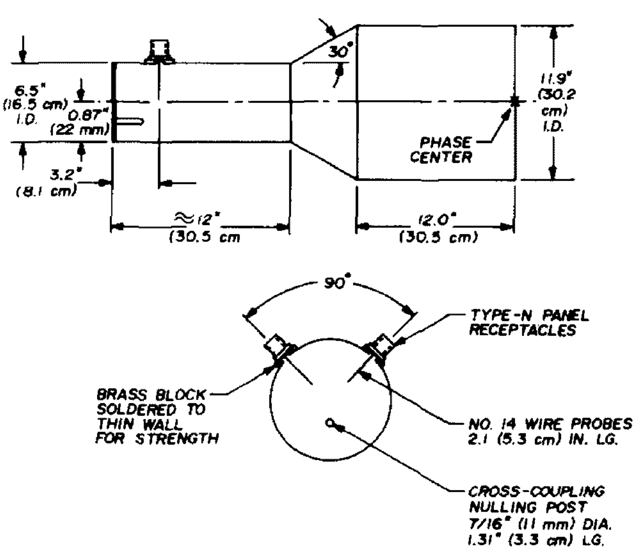


fig. 7. Dual-mode feed antenna for 1296 MHz. Material for the round sections is 1/32-inch (0.8-mm) brass, rolled and butt joint seamed. The conical section is soldered on with many small angle tabs. Probes are provided at right angles to facilitate circular polarization. Adjust probe lengths for best swr on 50-ohm feedline. The nulling post should be mounted in a radially slotted hole to permit radial adjustment of position for minimum cross coupling between probes.

larly polarized feed for 1296 MHz it is necessary to have an additional frame smaller in size and about 24 inches (61 cm) behind the first frame. Support members for this small frame are added between frames and side struts attached to the tripod legs with riveted straps for lateral support. If axial twisting of the feed support assembly is objectionable, it can be minimized by the addition of a rolled ring approximately 48 inches (1.2 meters) in diameter and fastened to the tripod legs in front of the feed support frame.

For 432 MHz the feed system shown in **fig. 7** may be mounted on long studs in front of the first frame so that the ground plane is located at the focal point of the reflector. This can be done with the 1296-MHz feed horn in place, but not in use, by mounting the ground plane just far enough in front of the horn to permit the coaxial feed cables to clear. The dual-mode horn is mounted inside the frames by means of short studs or brackets. It is important that screws used to fasten the horn should always be arranged with their heads on the inside of the horn wall.

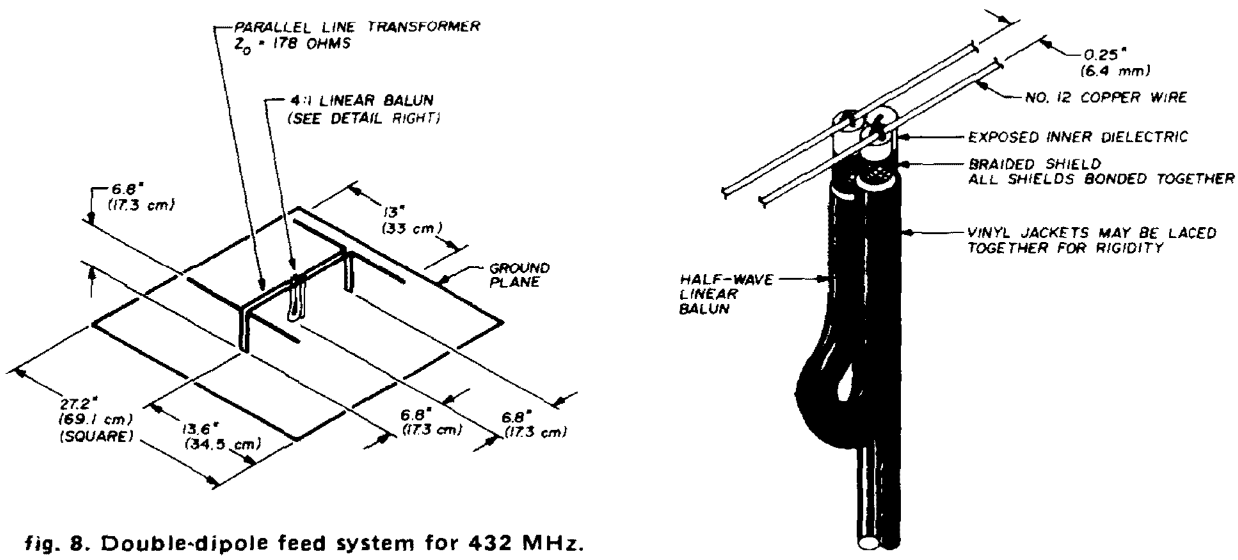


***fig. 8.* Double-dipole feed system for 432 MHz.**

counter weight

If the complete antenna is to be placed on an elevation-over-azimuth mount, the following construction will provide for counter balancing and extra stiffness. The counterweight arms are made of channel steel or aluminum beams spaced about 18-inches (46-cm) apart and cross braced in the manner shown in **fig. 6**. Attachment to the reflector hub is by means of two angle brackets which are bolted to the back of the hub. It is most helpful to place the elevation axis in the plane of one side of the support tower and as close to the hub as the support tower will permit. In this way the counter weight arms will lie parallel with the tower when the antenna is in the stowed position, beam aimed straight up. This is the position of least wind resistance and, by securing the counterweight arms to the tower, the reflector antenna and tower can survive severe weather conditions. It is also the best symmetrically balanced position for ice and snow loading.

final rigging

The final rigging involves turn-buckled guy wires from the end of each main rib back to the ends of the counterweight arms on their respective sides. The registration or positioning of the main ribs in the reflector with respect to the tower mount should be such that the bottom two ribs straddle the tower. In this way all rib ends may be guyed to the counterweight arms. If the bottom rib were aligned vertically it would not be possible

to guy it due to interference from the tower. All guys should be carefully trimmed since they essentially hold the reflector and counterweight arms in rigid assembly. Three additional guys may be added from the rib ends between the tripod feed support legs to the feed-support frame.

The counterweights are simply large, heavy bars U-bolted to the inside of the counterweight arms so that they may be slid along the arms for adjusting balance. Alternatively, the bars may be replaced by iron pipe sections which have heavy concrete weights permanently attached at one end.

summary

Construction techniques for relatively large paraboloidal reflector antennas have been given with the hope that those amateurs who are interested in high gain antennas for uhf will find comfort and encouragement in their endeavors.

Since feedline losses are significant at uhf, it is recommended that the preamplifier and mixer be located at the feed with only i-f and dc cables dressed down a leg of the tripod and around to the mount axis where a slack loop will permit rotation without cable joints. For strictly moonbounce communications, the support tower for the reflector need be no higher than the reflector radius, provided that the foreground clearance (trees, buildings, etc.) is good in the azimuthal angular directions of the rising and setting Moon.

It is also suggested that the final power amplifier and driver be mounted in a suitable weatherproof enclosure on the counterweight arms so that a minimum of large low-loss coaxial cable can be used to connect the output to the feed. This arrangement has special merit because no rotating joints or flexing cable are required in the high power line but only in the low power line to the driver input where lossy flexible cable can be tolerated in a slack loop.

A note of caution for those who live in extreme cold areas where icing and high winds occur: It is not known how this antenna construction will survive these extremes of weather. It is advisable in areas where freezing temperatures occur to drill small (1/8-inch or 3-mm) diameter holes in all tubing where water may collect.

The 16-foot (4.9-meter) reflector antenna described here was a first prototype and has proven adequate in receiving the first EME signals in Australia on 1296 MHz from W2NFA. Actually, the first signal ever heard on this antenna was W2NFA via EME!

acknowledgement

A project of this magnitude must require enthusiasm, help and encouragement and I am happy to acknowledge the assistance of the following people. Mike Hooper, VK3CCX, for his help in designing and building the jigs and trusses and for installing the cables; Les Jenkins, VK3ZBJ, for his contribution with the feed and adjustment, the loan of a converter and for numerous tests of equipment; Noel Walker, VK2ZNS, for helping with the design and building of the mount and digging holes for tower foundations at 1:30 AM (the weather was too hot during the day); Roger Hord, VK2ZRH, Ray Beevers, VK3BRB, and many others for various efforts related to the fabrication of the reflector and its final installation. I am particularly indebted to Dick Turrin, W2IMU, for his boundless encouragement and technical advice and lastly to my wife, Margaret, and my family for their patience and cooperation.

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ham radio

simple antennas for satellite communications

How to adapt ground-plane antennas for practical communications through OSCAR 6

Dale W. Covington, K4GSX, 281 Vance Circle, NE, Marietta, Georgia 30060

If the most consistent communication via amateur satellites is desired, then beam antennas which constantly track satellite passes over the ground station are clearly in order. On the other hand, the current OSCAR series of satellites present dramatic proof that many contacts through the on-board repeater can be generated by *stations using relatively low gain, stationary* antennas. Granted some freedom from complicated antennas and tracking mechanisms, it becomes worthwhile to investigate design modifications of simple fixed antennas which maximize the operating time during an arbitrary pass. The following comments suggest a design change for ground-plane antennas that makes them attractive for modestly equipped stations involved in satellite communications.

The basic concept is quite simple. Slightly tilting the vertical radiator tends to optimize the transfer of energy between a satellite in a circular orbit and a ground-plane antenna for any point locating the satellite in the visible sky. The design objectives and the theoretical basis underlying the modification are reviewed below. Practical tips on matching and constructing the modified antenna are also discussed.

design objectives

The following four characteristics pro-

vide a useful (but not exhaustive) set of design objectives for fixed antennas used for satellite work:

1. A vertical plane pattern which remains constant along any azimuthal bearing in the horizontal plane.

2. A vertical plane pattern which provides a smooth increase in radiation as the vertical angle of elevation, measured between the horizon and the satellite, decreases. An increase of 11 dB between the overhead and horizontal field strengths is appropriate for OSCAR satellites at an altitude of approximately 910 miles.[1]

3. Relative independence from ground effects. This permits installation of the antenna in the clear where nearby trees, buildings, etc., exert minimum screening when the satellite nears the horizon.

4. Simplicity in mechanical construction and electrical matching.

The selection of the first two items involves certain simplifying assumptions: the satellite is in a nonsynchronous-circular orbit, signal perturbations caused by interactions between the radiated wave and the ionosphere are negligible, and the satellite antenna is always favorably oriented with respect to the fixed antenna at the ground station.

theoretical basis

Which of the design objectives are met by a quarter-wavelength vertical erected adjacent to the ground? No serious problems concerning point 4 are encountered by this antenna, and the theoretical pattern of the isolated vertical completely satisfies point 1. Over perfect ground the radiation is maximum along the horizon which is convenient for DX work through the satellite. However, a depressing null develops in the vertical plane pattern for elevation angles near the zenith. Thus signals will be degraded during the overhead portions of satellite passes. The antenna fails on point 3. Two simple alterations remove these drawbacks.

It has been shown elsewhere that tilting the vertical element away from the normal fills in the overhead null without materially changing the other pattern characteristics.[2] A vertical plane pattern for a tilt of 45 degrees is drawn in **fig. 1A**. The depression at the zenith deepens and field strengths become less dependent upon the azimuthal bearing as the tilt

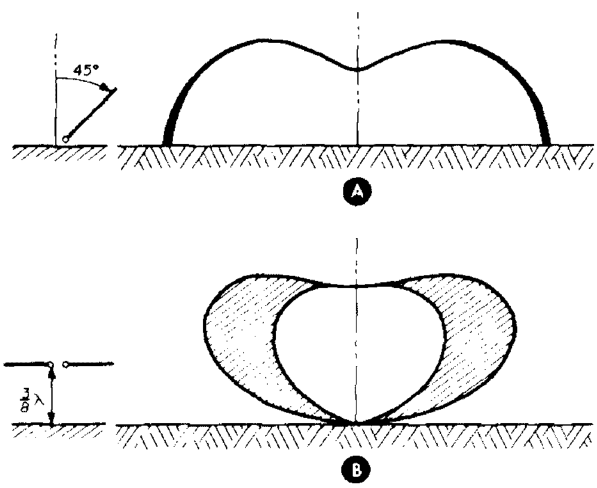


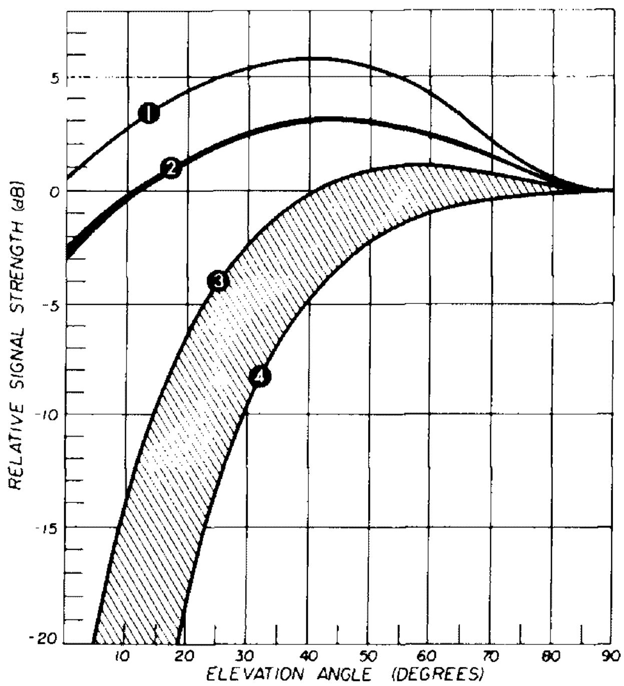
fig. 1. Vertical plane patterns for quarter-wave vertical tilted away from the normal by an angle of 45 degrees (A) and horizontal dipole at a height of 3/8 wavelength (B). The patterns for an arbitrary azimuth bearing fall within the width of the solid curve for the tilted vertical (A) and within the shaded region for the horizontal dipole (B). Perfect ground is assumed for both cases.

angle decreases. The width of the curve in **fig. 1A** reflects the extremes in vertical plane patterns as one circles a vertical tilted at 45 degrees. Since the variation is less than 0.5 dB, this trade-off is easily accommodated. Another trade-off involves the lower input impedances of the tilted vertical. As shown below, it is an easy matter to resolve this problem with a quarter-wavelength coaxial transformer.

The second change is to simulate actual ground with a plane of quarter-wavelength radials. This frees the antenna from the earlier criticism regarding point 3. Naturally the limited size of the elevated ground plane will slightly alter the pattern of **fig. 1A**. Basically, somewhat more radiation occurs at higher

angles and somewhat less energy is radiated along the horizontal direction. The net result of these alterations is a tilted, vertical, ground-plane antenna which adequately meets each of the four design goals listed above.

The advantages of this antenna can be brought into sharper focus by briefly examining the characteristics of horizontal half-wave dipoles using the same performance criteria. **Fig. 1B** gives the verti-



Curve 1 Vertical tilted 30 degrees from normal
Curve 2 Vertical tilted 45 degrees from normal
Curve 3 Horizontal dipole broadside to satellite
Curve 4 Horizontal dipole endfire to satellite

fig. 2. Relative signal strength versus angle of elevation between the satellite and the horizon. The curves can be used to estimate the effectiveness of a given antenna as the elevation angle changes during a satellite pass. A reference level of 0 dB was arbitrarily selected at the zenith point for each antenna. In curves 3 and 4 the dipole is assumed to be 3/8-wavelength above ground. The satellite is in a circular orbit at an altitude of 910 miles.

cal plane pattern for a horizontal antenna at a height of 3/8 wavelength. This height is close to optimum because it offers the greatest freedom from the undesirable lobe effects in the vertical-plane pattern associated with higher antennas while still maintaining a greater percentage of radiation nearer the horizon than can be supplied by lower antennas.

The outer and inner boundaries of the dipole pattern in **fig. 1B** define the broadside and endfire conditions, respectively. It is immediately obvious that the radiation from the dipole is sensitive to azimuthal bearing. Moreover, horizontal antennas at any height always show poor radiation along the horizontal plane. In practical terms at 10 meters, the average city-dweller will find that the effects of real earth and screening likely to be present for an antenna located only 12-feet above ground combine to further reduce antenna performance at low elevation angles. Thus, the single horizontal dipole fails to satisfy the first three points of the design criteria. Azimuthal omnidirectionality can be improved by connecting a second dipole at right angles to the first in a turnstile configuration.[3] Points 2 and 3 still remain inadequately satisfied, however.

A graphical summary of the theoretical performance of a tilted vertical and horizontal dipole is presented in **fig. 2.** The curves of relative signal strength incorporate not only the variation in radiated field strength of the ground station antennas shown in **fig. 1** but also the varying distance separating the antenna and the satellite as elevation angle changes during a pass. It is convenient to arbitrarily normalize all curves to zero dB at an elevation angle of 90 degrees. Therefore, comparisons between different antennas should be restricted to relative changes in signal strength for corresponding changes in elevation angle. **Fig. 2** indicates that a vertical tilted 30 to 45 degrees provides a response which varies less than 6.0 dB, even for an overhead pass. Tilt angles which fall outside this range cause antenna performance to deteriorate.

The curve widths provide a measure of the departure from azimuthal omnidirectional behavior for tilt angles of 30 and 45 degrees. A much larger variation in signal strength is observed for a satellite passing over a horizontal dipole located

3/8 wavelength above ground. The marked disparity between the broadside and endfire curves at low elevation angles illustrates the value of installing a rotator to keep the dipole broadside to the satellite when it is near the horizon.

While simple treatments of ideal antennas are helpful in formulating an overall design philosophy, the real antenna usually provides some deviation from the predicted behavior. For example, real

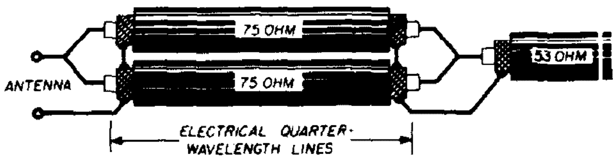


fig. 3. Quarter-wave matching transformer which matches antenna radiation resistances of the order of 25 ohms to the 50-ohm impedance level of popular coaxial lines.

ground, non-sinusoidal current flow, satellite spin and ionospheric polarization distortion can modify conclusions based on models which ignore there features. Low gain antennas aggravate matters because a host of conducting and insulating objects are illuminated in the immediate vicinity of the antenna. Precise pattern and impedance descriptions under such conditions are very elusive. The influence of these complications is usually not overwhelming, however, and practical antennas can be expected to approach the theoretical behavior deduced from ideal models.

matching transformer

As a vertical antenna is tilted away from the normal, the radiation resistance decreases from the theoretical value of 36.5 ohms computed for the perpendicular orientation. King's results for thin-wire, V-shaped antennas reveal that vertical tilts of 30 to 45 degrees yield radiation resistances of 29 to 21 ohms, respectively.[4] Although low-impedance coaxial cable (50-ohm class) can directly feed an antenna having a radiation resistance on the order of 25 ohms, a much better match is obtained by inserting a quarter-wavelength section of transmission line having an impedance equal to the geometric mean of the load and feedline impedances. This suggests a transformer impedance of about 36 ohms in the present application:

$$Z_T = \sqrt{Z_R \times Z_O} = \sqrt{25 \times 50} = 35.36 \text{ ohms}$$

Such a line is approximated by connecting two lengths of 75-ohm line in parallel as shown in **fig. 3**. The lines are an *electrical* quarter-wavelength long. Because the interior dielectric of conventional coaxial lines reduces the velocity of waves on the line, the *physical* length of the transformer section will be less than a quarter-wavelength in free space.

construction tips

Tilting the vertical radiator offers little complication of the detailed plans for building ground-plane antennas which abound in amateur periodicals and handbooks. Therefore, the following tips are simply offered to recall some well-known construction ideas using readily available parts.

table 1. Lengths of elements and matching transformers for tilted vertical ground-plane antennas where f is the frequency in MHz. Remember velocity factor, V, when computing the length of the coaxial quarter-wavelength transformer (0.66 for polyethylene cables, 0.79 for foam dielectric).

component	dimensions: high-frequency length (feet)	dimensions: vhf length (inches)
central radiator	$\frac{235}{f}$	$\frac{2800}{f}$
ground radials	$\frac{240}{f}$	$\frac{2860}{f}$
quarter-wave coaxial transformer	$\frac{246\ V}{f}$	$\frac{2950\ V}{f}$

Fig. 4 shows the details of a mounting assembly suitable for 10-meter ground-plane antennas. The four ground radials are cut from 10-foot sections of half-inch aluminum conduit using the lengths given in **table 1**. The remaining threaded end of

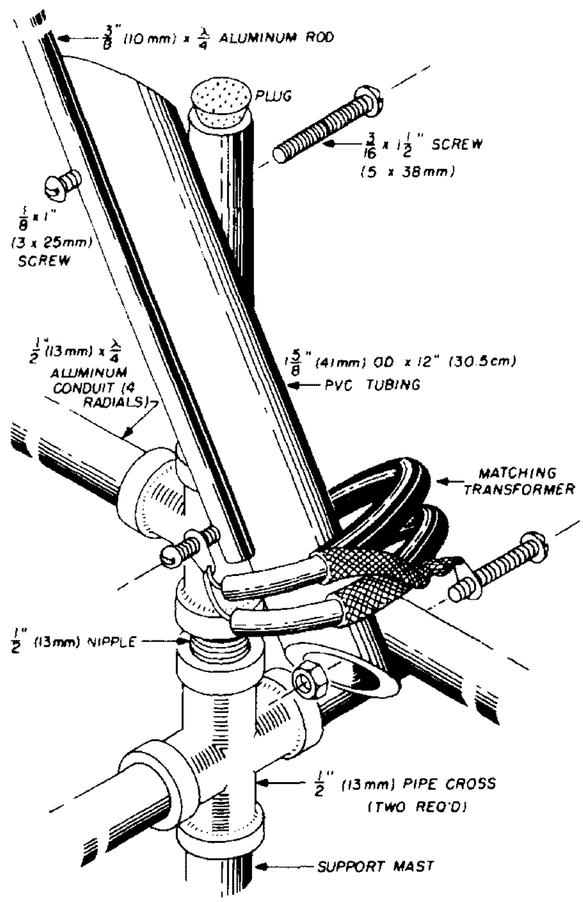


fig. 4. Isometric view of a partially assembled 10-meter ground-plane antenna. The whip and PVC tube are tilted with respect to the short upright section of conduit by an angle of approximately 30 degrees. While the antenna could be fed with a single-line of low-impedance coax, a better match is obtained by inserting a matching transformer constructed from two lengths of 75-ohm coax connected in parallel (see fig. 3). Table 1 lists the lengths of the tilted radiator and the ground-plane radials.

each radial is screwed into a half-inch pipe cross. A short upright section of conduit serves to anchor one end of the PVC water pipe used as an insulator for the center element. The opposite end of the PVC tube is fastened to one of the radials. As indicated in **fig. 4**, the PVC tube is cut at an angle to permit easy access to the nuts and washers securing the bolts which fasten the PVC tube to the mount and the whip to the PVC tube. The whip is formed from 3/8-inch aluminum rod. The rod, PVC tube and pipe fittings are stock items in most hardware stores. Aluminum conduit can be obtained from local electrical suppliers or contractors.

Lengths for the various elements and the matching transformer are listed in table **1**. Exact whip and radial lengths for resonance depend upon the tilt angle and the relative element diameter with respect to the operating wavelength. The frequency response of these antennas is broad enough (3 percent bandwidth for vswr less than 1.5:1) that precise adjustment of the element lengths should not be necessary, however.

Since the pipe joints are painted with a metal primer to retard galvanic action, positive electrical contact is insured by installing wire straps near the threaded end of each radial. In the interest of clarity, these straps and the sheet metal screws which hold the straps in place are not shown in **fig. 4.** All electrical connections are sealed from the weather with silicon caulking compound.

The tilted vertical ground plane can also be used to advantage on the vhf links of OSCAR satellites. One of the simplest ways to construct these antennas is to use stiff wire or 1/8-inch welding rod for the elements and standard coaxial chassis receptacles (SO-239) for the base mounts. Lengths appropriate for vhf designs are also given in **table 1.**

observations

To some extent theoretical concepts and drawing board creations present an aura of make-believe. There is one central question at this stage that needs an answer. Does the modified ground plane deliver real performance during actual satellite passes?

The 10-meter downlink signals of numerous OSCAR 6 passes were monitored on a tilted vertical ground plane (35-degree tilt, height of 3/4 wavelength) and a half-wave dipole (height, 3/8 wavelength). Both antennas were well matched with vswr below 1.5:1 in the downlink passband. For elevation angles below 30 degrees, the modified ground plane was clearly superior.

Switching to the ground plane would often provide Q5 copy of signals that were only Q2 to Q3 on the dipole. This

improvement is particularly significant since most of the operating time available for OSCAR satellites occurs for elevation angles below 30 degrees. The dipole came on strongly at elevation angles near 90 degrees. Although signals remained good on the ground plane, the results of these tests indicated that the signals on the dipole were stronger by roughly 4 dB when the satellite was directly overhead. The most consistent copy throughout any pass was always obtained on the ground plane.

Using the projected capabilities of AMSAT-OSCAR-B satellites, a few remarks are in order concerning applications for modified ground planes over the vhf links. Mechanical tracking schemes can be eliminated with these antennas if a transmitter power of approximately 200 watts is available on the 432-MHz uplink or 80 watts on the 145-MHz uplink. The additional gain from beam antennas will be useful for downlink reception in these two bands, but under optimum conditions signals should be copied with low-gain antennas such as dipoles and ground planes.

summary

This pretty well wraps up the case for adapting ground-plane antennas to satellite communications. Slightly tilting the vertical radiator yields a stationary antenna that is useful for any satellite pass. The modified ground plane is a simple way of getting the job done, but you don't have the extra performance of a tracking beam. You also don't have to pay for multi-elements, two rotators, tracking charts, a mini-computer or extra arms.

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ham radio

vertical antenna ground systems

How to design practical, efficient radial ground systems for vertical antennas

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

This is the third article of a series about vertical antennas. The objective is to learn about the characteristics of such antennas so that you can intelligently select an antenna height for your own station. The first article[1] showed that for a short antenna (h = 0.1λ), bandwidth was small (50 kHz at 3.8 MHz), and matching-network coil losses were high (about 100 watts). The second article[2] showed that the radiation pattern is affected very little by height (in the range from h = 0.1 to 0.25λ).

In this article I will show the effects of earth losses and the radial system upon the ability of a vertical antenna to radiate efficiently. The radial system consists of a number of radials, of a certain length, using a given wire size, and buried in earth having a finite conductivity. The radiating efficiency of the antenna depends upon the radial system, earth conductivity and antenna height.

ground systems

My analysis of earth losses and the effectiveness of radial systems uses two approaches, one using a theoretical model for earth losses, and one using actual experimental data. The experimental results give field strength at 3 MHz at one mile from the antenna for various radial conditions. While the theoretical model does not allow calculation of exact values of earth losses in watts, it does provide insight into important situations not covered by experimental data, such as losses for poor earth and the distribution of earth losses near the antenna for various conditions.

Both of these methods are from Brown's 1937 article,[3] and they are applied to the amateur situation used in my previous articles for a 30- to 60-foot (9.1- to 18.3-meter) vertical antenna at 3.8 MHz. The results, however, are given in terms of antenna electrical height and radial electrical length to allow you to design vertical antennas and radial systems for other amateur bands as well.

Since an antenna and a radial system form a closed electrical circuit, there are return currents flowing in the radial wires and in the earth itself near the antenna. Since there are so many variables involved I wrote a computer program to calculate these currents for a variety of factors, listed in table 1. For each antenna height the correct antenna base resistance and

matching-coil rf resistance was included. This was done for a transmitter power of 600 watts output at 3.8 MHz.

The 600 watts is divided between radiated power from the antenna, and losses in the earth, matching coil and radial wires. Ideally, of course, we would *like the radiated* power to be 600 watts and the losses to be zero. Interestingly enough, there *are* combinations when this *almost* happens.

To simplify the analysis I tried to eliminate unimportant factors, and the most likely one was wire size. My analysis showed that wire size is important only when a few radials (eight or less) are used, when the antenna is short (less than 0.15λ), and for poor earth (σ = .00002 mhos/cm^3). For this case, going from number-18 wire (1-mm diameter) to number-8 wire (3.3-mm diameter) will increase the radiated power up to as much as 125%. However, by using 15 or more radials, wire size becomes relatively unimportant and the improvement then, by going from number-18 to number-8, never exceeds 8%. On this basis I eliminated wire size as a factor in the remaining analysis.

The power lost in the radials is insignificant. Wire conductivity is so huge compared to earth conductivity that you can use copper or iron wire. Number-18 iron electric fence wire (1-mm diameter) costs only $7.10 for 2640 feet, but number-16 copper wire (1.3-mm diameter) would be better for a permanent installation as it would not corrode and disintegrate as quickly as iron wire.

radial system design

To design the radial system you need to know how many radials to use and how long they should be. A look at how the radial currents drop off as you go further from the antenna provides the necessary information. This is illustrated in **fig. 1**, which shows radial current (I_W amperes) versus distance from the antenna in fractions of a wavelength (x/λ), and also versus the number of radials for both poor and good earth. When the radial current drops to a low value the radials are no longer effective. At this distance from the antenna base most of the current has already entered the earth, and longer radials are unnecessary.

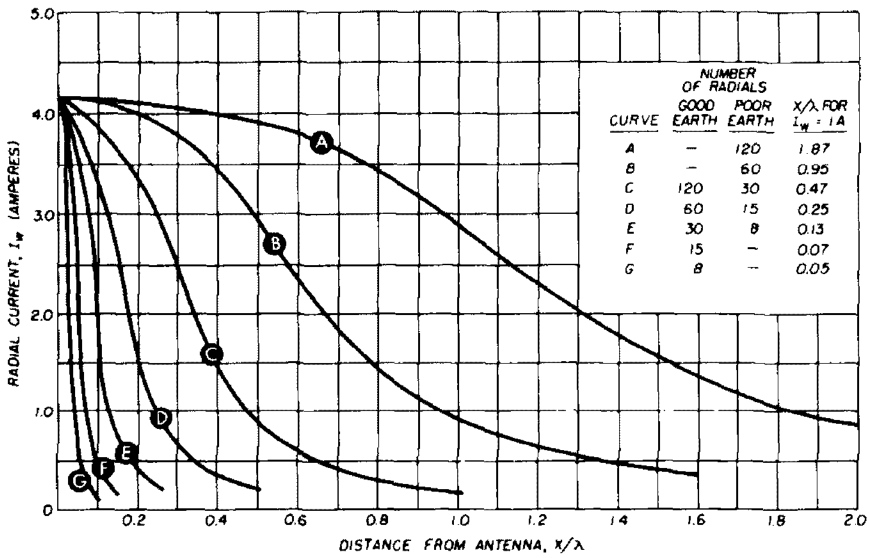

fig. 1. Radial current vs distance from antenna for different numbers of radials in poor and good earth.

You can see from **fig. 1** that long radials should be used with a large number of radials, and vice versa. Therefore, this situation for vertical antenna radiation is somewhat different from that of obtaining a low ground resistance where you can trade-off number of radials for length of radials. The proper radial length depends upon the number used, local

earth conductivity, how low you want the radial current to be at the end of the radials, and how much real estate you can use. These results are summarized in table **2** for a radial current at the ends of the radial wires of 1 ampere.

Table **2** shows that, if you have poor earth, the radials should be two to four times longer than for conditions of good earth. There is, of course, no harm in using longer radials, and you can use **fig. 1** as a guide for your own location.

There are several criteria for deciding how long radials should be. The one used in table **2** is for radial current, I_W, to drop to 1 ampere. Other criteria are given later in the discussion about **figs. 2** and **3**.

The question of how deep to bury radials often comes up. Here the earth losses are calculated down to the skin depth, which at 3.8 MHz is several feet (about 1 meter). Both the theoretical model and the experimental data show that it doesn't matter if the radials are buried a few inches. This is better than tripping over a maze of wires strung out on top of the ground (your wife will be happier, too).

I next calculated earth losses for various radial conditions, for both poor and good earth. The losses are in watts/meter of distance, down to the skin depth, at various distances from the antenna base, and are shown in **fig. 2**. Study of this figure shows:

table 2. Radial length to reduce radial current to 1 ampere (assuming 600 watts at 3.8 MHz).

number of radials	distance from antenna, poor earth (wavelength)	distance from antenna, good earth (wavelength)
8	0.13	0.05
15	0.25	0.07
30	0.47	0.13
60	0.95	0.25
120	1.87	0.47

1. Maximum values for earth losses are high when only a few radials are used.

2. When only a few radials are used the losses peak close to the antenna base (within 0.1λ for eight radials).

table 1. System design factors considered for calculating return currents for a vertical antenna and radial ground system.

Antenna heights, feet	30, 40, 50, 60 (9.1, 12.2, 15.2, 18.3 meters)
Number of radials	8, 15, 30, 60, 120
Radial wire size	no. 8, no. 18
Radial lengths	0.05 to 2.0 wavelength
Earth conductivity	0.00002 mhos/cm^3 (poor earth)
	0.0004 mhos/cm^3 (good earth)

3. The total amount of power lost in the earth is related to the area under each curve. It shows that, for the same number of radials and antenna height, considerably greater earth losses occur in poor earth than in good earth.

4. Higher values of losses occur closer to the antenna base for good earth as compared to poor earth, other conditions being the same.

5. When many radials are used the losses are not greatly different for 30- or 60-foot (9.1- or 18.3-meter) antennas. When only eight radials are used, then losses are greatly reduced by using a 60-foot (18.3-meter) antenna instead of a 30-foot (9.1-meter) one.

Coil losses were similar to those given in the previous article.[1] These losses, and some other previous data, are summarized in **table 3**. The coil losses are somewhat less when earth losses are considered than they were when earth losses were not considered. Part of the 600-watt transmitter output power now contributes to earth losses, so that coil losses and radiated power are lower than they would be otherwise.

Some of the experimental results from reference 3 that apply to our situation are shown in **fig. 3**. This shows radiated

power as indicated by field strength, E, at ground level at one mile in millivolts/-meter, versus number and length of radials and antenna height. These results appear to be for conditions of good earth conductivity, where coil losses were not included. Comparison curves for the theoretical maximum possible values for E are included, and these were probably derived from measurements of antenna base current. A study of **fig. 3** reveals several interesting facts:

1. When using a 30-foot (9.1-meter) antenna and two radials 0.137λ long, the distant field strength is only about 50% of theoretical maximum. This is for good earth and does not take coil losses into account. They will be considered later.

2. Using a 60-foot (18.3-meter) antenna and 113 radials, 0.41λ long, gives results nearly as good as the theoretical maximum.

3. When only two radials are used they need be only 0.1λ long.

4. When 15 to 30 radials are used with a 30-foot (9.1-meter) antenna 0.274λ is long enough for the radials.

5. Other situations may be deduced from **fig. 3**, and from methods of design discussed later.

local earth conditions

We need to define what poor and good earth is. I have used a conductivity of σ = .00002 mhos/cm³ for poor earth, and σ = .0004 mhos/cm³ for good earth. You should be curious by now to know what the conductivity is for your local area. This can be found in Jordan,[4] page 638, or in reference **5**.

It is beyond the scope of this article to relate radiated power to the field strength at radiation angles above the horizontal. My recent article[2] gave the pattern for radiation angles of DX interest for the case of no ground losses. When *ground* losses are considered, as must be done in the *real* case, the low-angle radiation will depend on the earth conductivity. See Jordan,[4] page 641, for more information on this. Hovever, whatever the earth conductivity, the less your losses are by use of a good radial system, the more power you will radiate.

choosing antenna height

Now we should go back and think about our original objective: Choosing an antenna height. The yardsticks that should influence the decision are first,

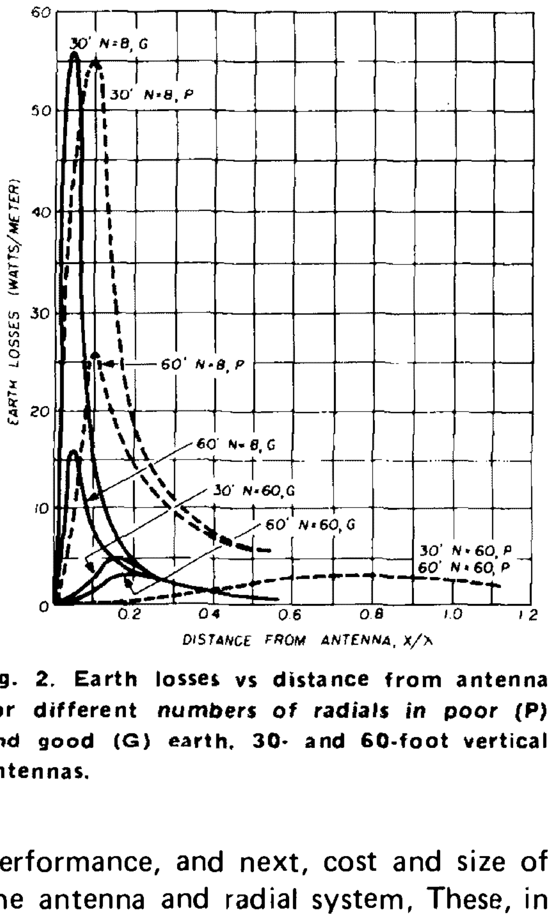

fig. 2. Earth losses vs distance from antenna for different *numbers of radials in poor* (P) and good (G) earth. 30- and 60-foot vertical antennas.

performance, and next, cost and size of the antenna and radial system, These, in turn, will depend, as we have seen, on the number of *radials used, the length of the* radials, and the local earth conductivity. Most of the results have been summarized in **table 3.** Earth losses and radial systems were not included in **table 3** because there are so many possible combinations. However, I'll work out some examples *which will show you how to design for* your own situation.

example 1. Antenna height, HA = 60 feet (18.3 meters) good earth, number of radials, n = 113, radial length, φ = 0.41λ. **Fig. 3** shows that the distant field strength, E, is about 99% of theoretical. In all of these examples you can relate E

to the radiated power, P, since E is proportional to $\sqrt{P}$. Therefore:

$$\frac{E}{E_{TH}} = \sqrt{\frac{P}{600}}$$

Where E_{TH} is the theoretical maximum, or 100%. In this case,

$$\frac{E}{E_{TH}} = \frac{99\%}{100\%} = 0.99 = \sqrt{\frac{P}{600}}$$

so P = 594 watts. Since coil losses are not yet included, the earth losses are 600 − P = 600 − 594 = 6 watts, and are low due to the good radial system. Total losses = coil losses + earth losses = 1 + 6 = 7 watts. The field strength, E, using this total loss is still about 99% of theoretical.

example 2. Antenna height, HA = 30 feet (9.1 meters), poor earth, n = 60, φ = 0.25λ. **Fig. 3** is for good earth only. **Fig. 1** provides information on how to relate poor and good earth radial systems. To get the same results in poor earth that you would in good earth you should make the radials from 2 to 4 times longer for the poor earth case. If you limit the length as we have in this example, then the results will be inferior for the poor earth case. Refer to **fig. 1**.

For 60 radials and x/λ of 0.25, for poor earth (curve B), I_W = 3.95 amps (which also indicates a poor radial system). For this same value of radial current, I_W, and for good earth (curve D), x/λ = 0.1. Therefore, use x/λ = 0.1 in **fig. 3** which shows that E, before coil losses, is 70% of maximum. The corresponding earth loss, calculated as in example **1**, is 306 watts. Therefore, the total loss is earth loss plus coil loss, or 306 plus 80 watts, for a total of 386 watts. The resulting field strength, E, is 60% of maximum.

An example that you can practice on is to verify that E is only 22% of maximum for a 30-foot (9.1-meter) antenna, with two radials, each 0.1λ long in good earth.

As a final example, consider the difference in performance for two antenna heights for the same radial system and for good earth. Use n = 30, x/λ = 0.25, good earth, and for a 30- and 60-foot (9.1- and 18.3-meter) antenna. **Fig. 3** shows that before coil losses, an E of 82% for the 30-foot (9.1-meter) antenna, and an E of 84% for the 60-foot (18.3-meter) anten-

table 3. Vertical antenna coil losses and bandwidths for various antenna heights. Earth and radial losses are not included because of the many possible variations.

antenna height (feet)	(wavelength)[1]	coil losses (watts) good earth	poor earth	bandwidth (kHz)[2]
30 (9.1 m)	0.12	120	80	50
40 (12.2m)	0.16	45	35	100
50 (15.2m)	0.20	20	12	180
60 (18.3m)	0.24	1	1	320

1. Electrical height at 3.8 MHz.
2. No earth losses and 3.8-MHz matching network.

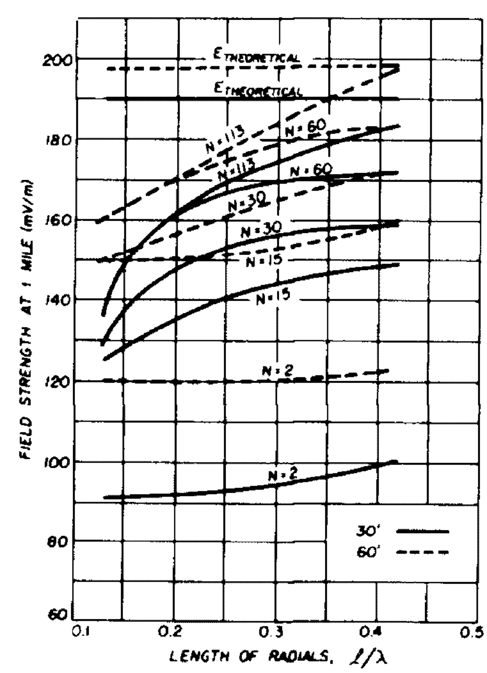

fig. 3. Experimental data from reference 3 showing field strength at one mile vs length of radials at 3 MHz for 30- and 60-foot antennas. This data assumes good earth and no coil losses.

na. When coil losses are included, the E for the 30-foot (9.1-meter) antenna is 68% of maximum, and the E for a 60-foot (18.3-meter) antenna is still 84% of maximum. Results for a 40- or 50-foot (12.2- or 15.2-meter) antenna fall between these.

conclusion

It is obvious that there is no simple answer to the question of what height antenna and what radial system to use. I began all of this work not knowing exactly what to expect. I had expected that short antennas (say 0.1λ high) would have shown up better than they have. It appears to me that a vertical antenna 0.2λ to 0.25λ high offers some significant advantages over one 0.1λ high: Essentially no coil losses, reasonable bandwidth (200 to 300 kHz at 3.8 MHz) and distant field strengths of 85% to 95% of theoretical with a reasonable radial system. There does not appear to be any advantage in going to a height of over 0.25λ for use at DX sky-wave radiation angles and, in fact, there may be some disadvantages when you consider the radiation pattern (see reference 2).

An important lesson from these studies is that earth conductivity is a very important factor, so that for poor earth long radials should be used. We have also seen that wire size is not generally important, and that it is permissable to bury the radials.

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measurement techniques

for antennas and transmission lines

How to use homebrew test equipment to measure impedance and resonance characteristics of antennas and transmission lines

John R. True, W4OQ, 10322 Georgetown Pike, Great Falls, Virginia 22066

Two **basic instruments** for the measurement of antenna and transmission line parameters that can be profitably used by the radio amateur are the radio-frequency impedance bridge and the spectrum analyzer. The radio-frequency impedance bridge is a most useful and informative piece of equipment with which both resistive and reactive components of impedance can be determined as separate values. The spectrum analyzer will display the resonant frequency of an antenna or a section of transmission line.

the rf impedance bridge

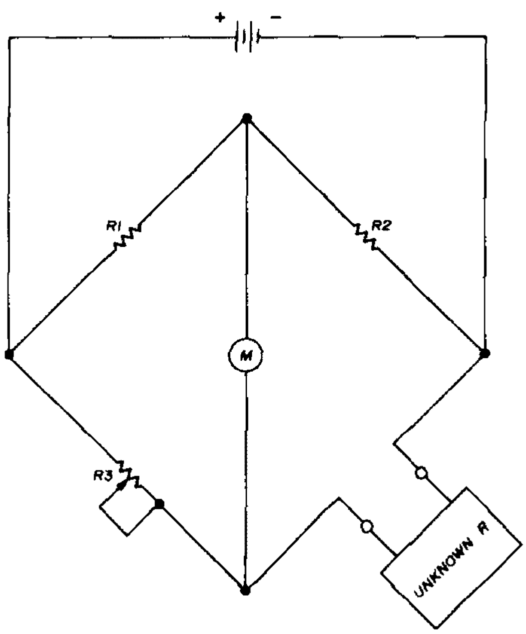
The standard circuit for a bridge is shown in **fig. 1A**. In this form, only resistive measurements can be made. But this bridge, while not useful for rf measurements, illustrates the basic bridge principle. Note that the excitation voltage is injected into two opposite arms of the bridge. When the bridge is balanced (R3 = $R_{unknown}$), there will be no voltage across the detector.

The bridge circuit shown in **fig. 1B** is equipped with provisions for ac excitation and the arms are arranged so that the resistive component is indicated separately from the reactive component. An excellent design for use on the amateur high-frequency bands was described in a recent issue of *QST.*[1] For use in the radio-frequency spectrum, the source generator may be any signal generator having a frequency range that meets the required

range. The detector can be any frequency-selective voltmeter such as a radio receiver's S-meter. When the bridge is balanced the voltage at the detector arms will then approach zero and a null will be detected. At this point the readings of the two variable arms will be representative of the unknown's internal impedance. Practical uses for this instrument include, but are not limited to, measurements of antenna impedance, tuned-circuit impedance, matching-transformer impedance, as well as the characteristic impedance of a transmission line.

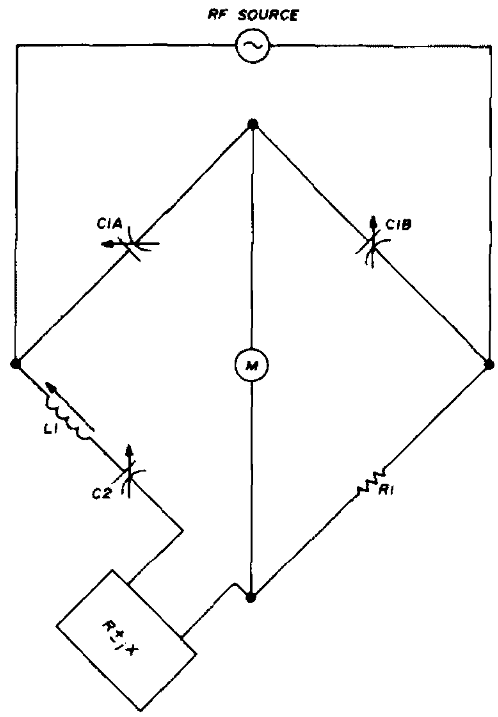
If the signal generator and detector are set to 7.00 MHz and the X dial indicates 210 when the bridge is balanced, the actual reactive ohms will be (210/7.00) = 30 ohms. Since the resistive dial is not frequency selective, the resistance dial indicator will not require such conversion.

If the indicated reactive component is inductive the result is written +j30. Con-



R1 = R2. METER NULLS WHEN R3 = $R_{UNKNOWN}$

A



DIFFERENTIAL CAPACITOR C1 CALIBRATED IN RESISTANCE. CAPACITOR C2 CALIBRATED IN REACTANCE. L1 ADJUSTED FOR NULL WITH 50-OHMS ACROSS UNKNOWN TERMINALS

B

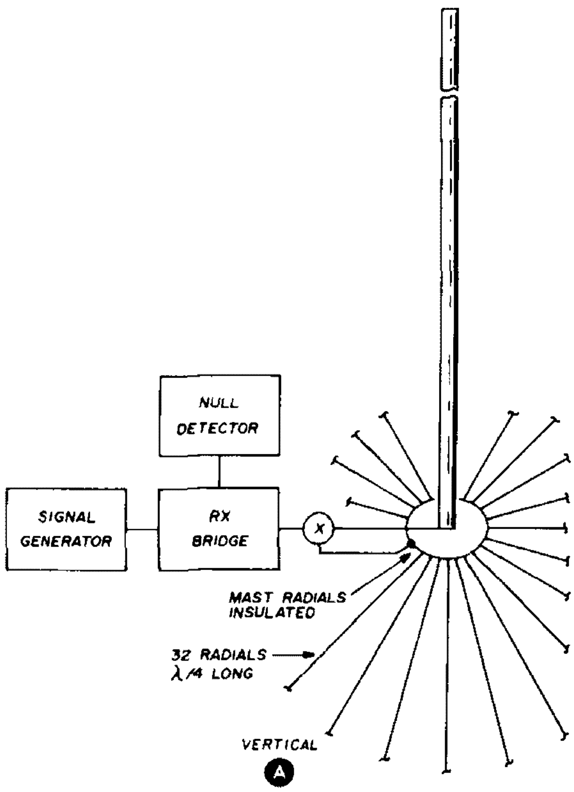
fig. 1. Basic balanced resistance bridge shown in (A) is not suitable for rf measurements. The RX impedance bridge shown in (B) indicates the resistive and reactive components of a complex impedance.

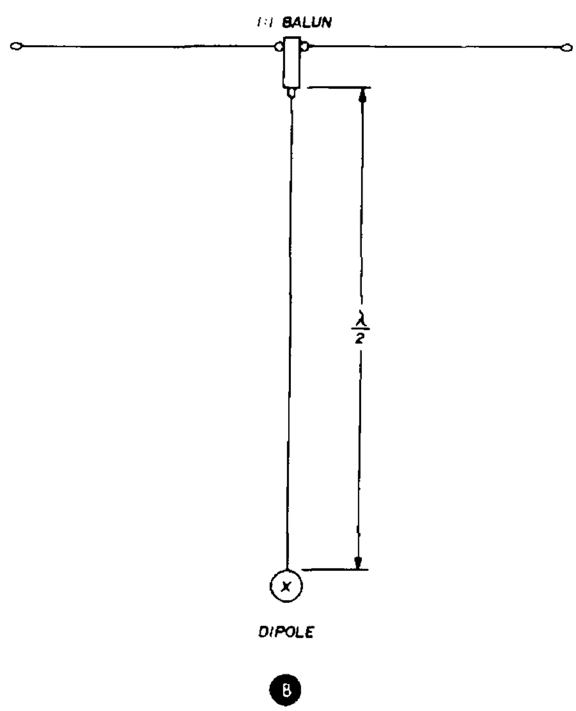
Setting up the instrument for measurement of an unbalanced system such as a vertical radiator is shown in fig. 2. Since most bridges are asymmetrical (unbalanced) instruments, they lend themselves to such measurement with ease. Readings on the X (reactance) dial must always be related to the frequency of calibration. For example, if the bridge was calibrated at 1 MHz, all indicated readings of the X dial must be divided by the frequency of measurement expressed in megahertz. An example will serve to clear up this point:

versely, if the reactive component is capacitive the result is written −j30. The physical representation of these values can be shown on rectangular coordinate graph paper (the +j30 value is plotted as 30 units *above* the zero axis (horizontal)— a −j30 value is plotted at 30 units *below* the zero axis. The resistive readings will be plotted to the right of the vertical axis in resistive units, as shown in fig. 3. An excellent article on the use of graphic solutions to impedance-matching problems may be found in reference 2.

balanced impedance measurements

The impedance of a dipole can be measured quite accurately by placing a 1:1 balun between the bridge and the





symmetrical radiator. This device converts the balanced impedance of the dipole to an unbalanced condition for measurement by the asymmetrical rf impedance bridge. The impedance can even be measured remotely, with good accuracy, by the use of an electrical half-wavelength transmission line (or multiple) and a balun, since the impedance seen at

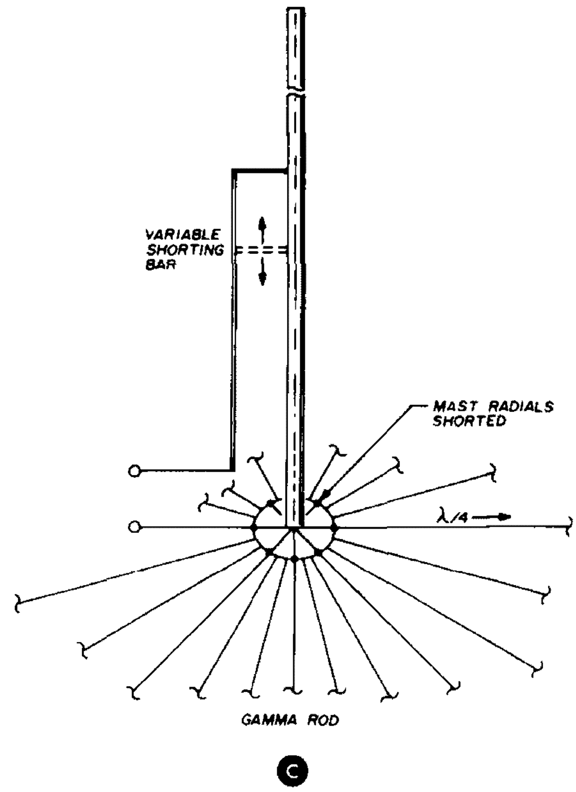


fig. 2. Using the RX bridge for impedance measurement of a vertical antenna (A), dipole antenna (B) and gamma-rod matching section (C).

the far end of a half-wavelength line is the same as that seen at the near end.

Odd wavelengths of line can also be used for remote impedance measurement, but the use of a device such as a Smith chart will be required to properly rotate an impedance value measured the correct number of electrical degrees, to obtain meaningful data. While this is a useful and practical method, the use of the Smith chart will not be covered in this article since several practical articles on its principles and use have appeared in recent amateur magazines.[3,4] How to measure electrical length of a transmission line will be covered in a later paragraph.

using the rf bridge

In practice, the use of the rf impedance bridge is quite simple. When connec-

ted to an unknown impedance as shown in fig. 4, The frequency of the signal generator is set to the desired frequency. The frequency of the detector is set to produce maximum indication of signal input. A "standard" is placed across the unknown terminals (in some rf bridges the terminals are shorted). When balanced, the bridge is ready for measurement of the unknown impedance. If the detector has no visual indicator (such as an S-meter), a pair of headphones can be effectively used to detect the null when the bridge arms are in balance. A little practice with this method can produce greater accuracy than a visual indicator, because the human ear can detect a deepter null than can a visual indicator.

When measuring the impedance of a tuned circuit, the bridge *will indicate* the correct operating parameters only if its normal voltages and current are present under certain circumstances. In many cases this power could damage the mea-

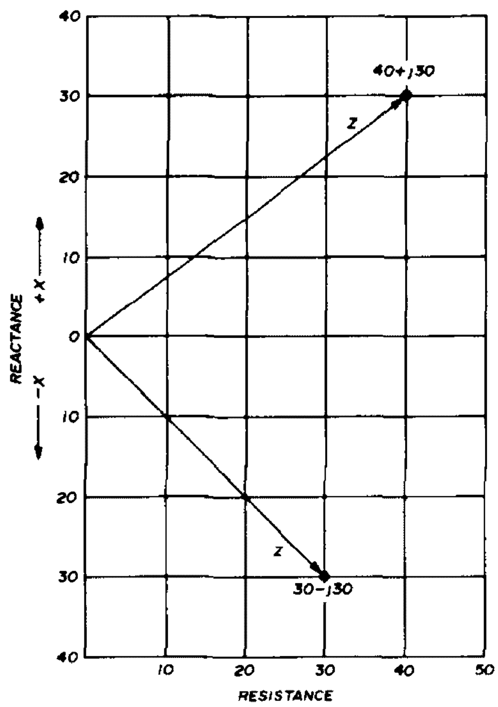


fig. 3. Plotting complex impedance values.

suring instruments so static substitution of such dynamic parameters must be made. One practical method is to substitute a resistor of the proper value for the dynamic value of resistance. First, determine the required resistance value by dividing the required operating voltage by the current, and put that value of resistance across the circuit to be measured

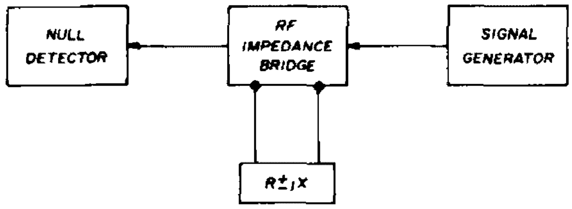


fig. 4. Typical rf impedance bridge setup.

without operating voltages and current. For example, assume you want to measure the input circuitry of a grounded-grid linear amplifier with a bifilar-wound filament-grid isolation choke plus tube and stray capacitances. From available data it may be determined that 400 mA of total dc filament current is required at 100 volts. Therefore, a 250-ohm resistor (100 volts/400 mA - 250 ohms) from filament to ground will simulate the dynamic desired operating condition and the bridge will indicate the proper load impedance shown by such circuitry (see fig. 5). Similar dynamic substitution can be made to the output circuit for proper plate-load design.

spectrum analysis

To some readers this may sound like a very sophisticated term. It is really very simple when its operation and technique is explained in basic terms. The equipment setup shown in fig. 6 is capable of spectrum analysis, although I prefer to call it the *sweep-null* technique as applied to the relationship of frequency vs electrical length measurement. It has many other applications, but for the purposes of this article I will stick to this basic use.

The sweep-frequency generator is simply an fm signal generator that is swept through a predetermined range of frequencies. Suppose a sweep of 4 to 5 MHz is desired. The sweep range is set to about 1 MHz and the center frequency is set to 4.5 MHz. This signal is then applied to the unknown through a detector and a

sample of the resultant is brought out to an oscilloscope for analysis. If the unknown impedance is frequency sensitive it will change impedance at various frequencies and thus load the detector accordingly. This will be evident as a varying voltage on the scope trace. To determine the exact frequency at which some parameter of the unknown impedance has a low or high value, a frequency marker which will appear on the scope trace at the appropriate point is injected into the detector.

If the unknown is a pure resistance, the horizontal trace of the scope will be a straight line with a vertical beat note pattern appearing on the trace which is

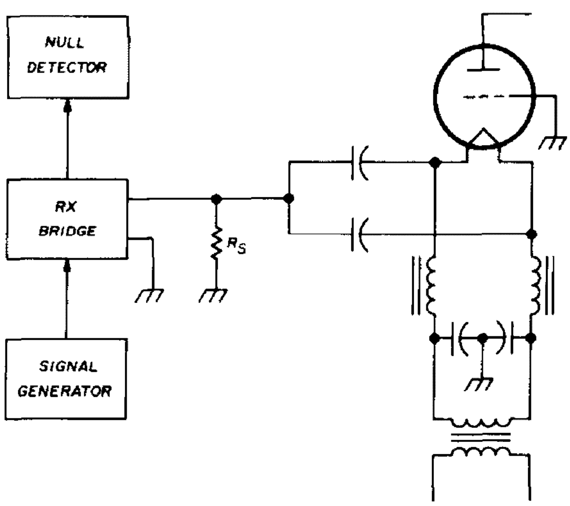


fig. 5. Method of substituting resistor Rs for dynamic resistance at the input of a grounded-grid cathode circuit.

dependent upon the frequency of the marker generator. However, if the load is a complex impedance somewhere within the sweep-frequency range, at this point the load will be frequency sensitive and will have a lower or higher ratio of voltage current. When this occurs, there will be a null or hump in the horizontal scope trace. If a wide frequency range is swept, there may be several repetitive nulls and humps in the trace which will coincide with lower or higher impedance points. If you want to determine the exact frequency of a hump or null, the marker is adjusted until the vertical mark is symmetrically centered.

By increasing the gain of the horizon-

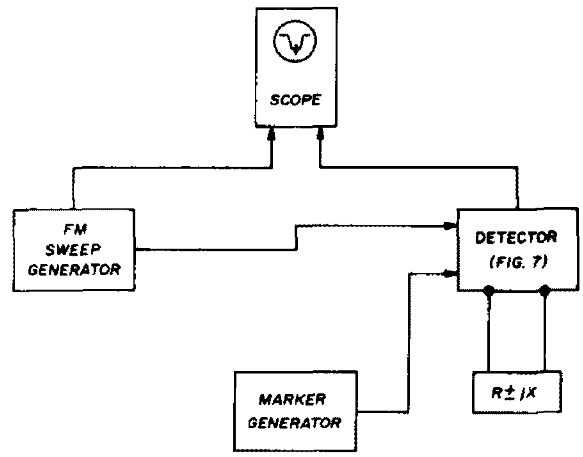


fig. 6. Sweep-null method of measuring resonance of complex loads.

tal and vertical scope amplifiers, a W-shaped trace for a null (or an M for a hump) similar to that shown in **fig. 8** will be formed and further symmetry can be achieved. The use of a horizontal graticule is often useful in judging symmetry. When the marker frequency is read it can be converted to electrical length by the use of standard formulas for propagation velocities in the various types of transmission line.

To illustrate a specific use of the sweep-null technique refer to **fig. 9**. This is a typical amateur setup using readily available equipment to solve a very practical problem: The measurement of a quarter- or half-wavelength (or multiple) of coaxial line.

To digress a moment, coaxial cable using solid polyethelene dielectric has a velocity constant of approximately 0.66. This means that this cable will propagate a signal at approximately two-thirds the

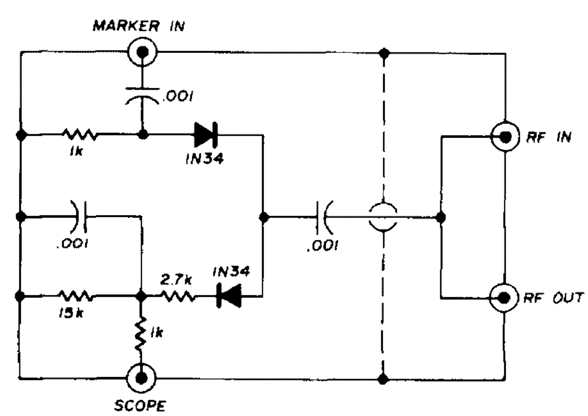


fig. 7. Circuit for the sweep-null detector used in the setup shown in fig. 6.

velocity that a signal will be propagated in free space. Another way of saying this is that it will take the same time for an rf signal to traverse 66-feet (20.1-meters) of that cable as would be required for the same signal to propagate 100-feet (30.5-meters) in free space. Foamed polyethelene dielectric has a velocity constant of about 0.80 so it will take about 80 feet (24.4 meters) of this cable to be electrically equal to 66-feet (20.1-meters) of solid dielectric polyethelene cable. The

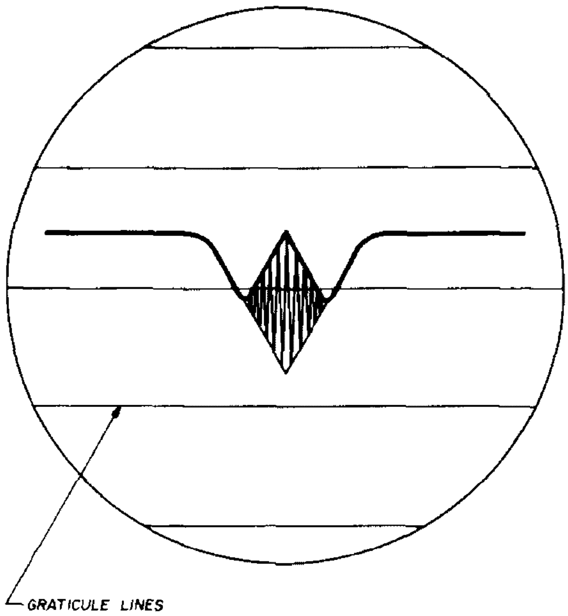


fig. 8. Oscilloscope pattern showing the W-shaped null pattern. Marker generator may be used to accurately determine the null frequency.

relationship between wavelength in free space and frequency in megahertz is given by:

$$\text{wavelength (feet)} = \frac{980}{f_{MHz}}$$

$$\text{wavelength (meters)} = \frac{300}{f_{MHz}}$$

Returning to the instrumentation shown in **fig. 9**, a 66-foot (20.1-meter) length of solid-dielectric coaxial cable, shorted at the far end, is connected to the detector. When swept over the frequency range from 3 to over 10 MHz, a null will be seen at 4.90 MHz (length = ½ wavelength). A hump will also appear at 7.35 MHz (3/4 wavelength) and another null at 9.80 MHz marks the full-wavelength

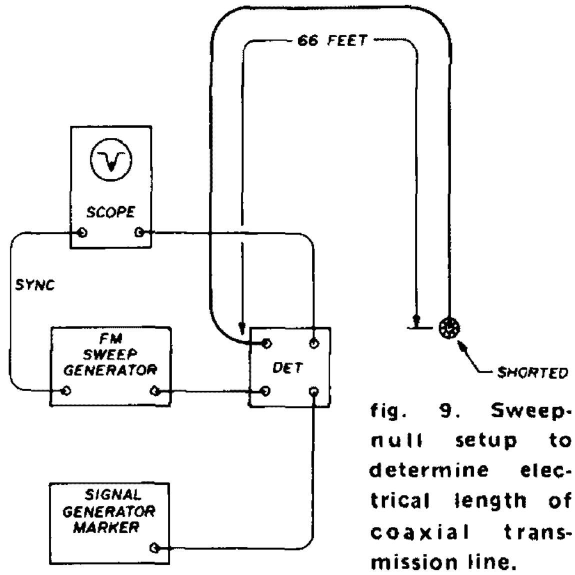


fig. 9. Sweep-null setup to determine electrical length of coaxial transmission line.

point. A practical note: The peaks on the scope trace are much less sharply defined than the nulls, since the *ratio* of change of voltage to current is more sharply defined at the low-impedance points (half-wavelength points). In practice this means that more accurate frequency determination can be made at the nulls than at the peaks.

As can be seen in fig. 10 the detector's inner conductor is connected to the

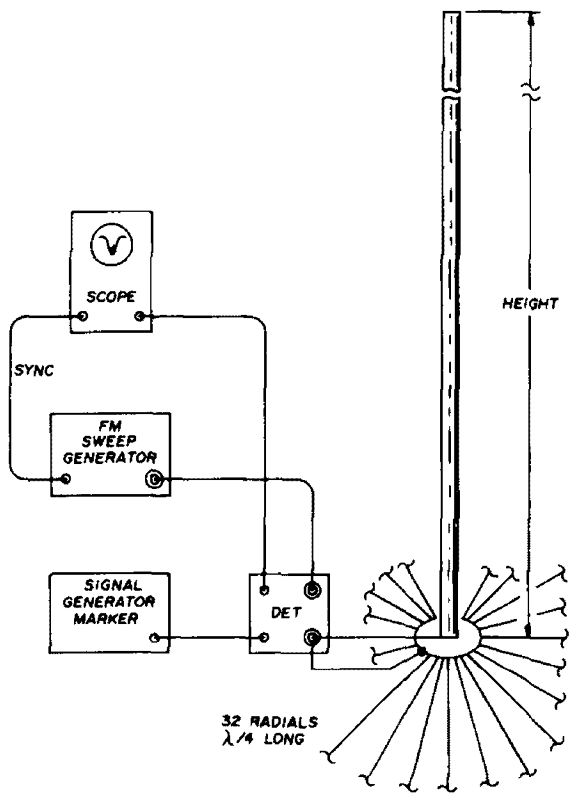


fig. 10. Sweep-null setup to determine electrical height of vertical radiators.

bottom of the vertical radiator (insulated from ground for this measurement) and the outer conductor is connected to the ground plane (*not* ground). As the sweep frequency goes through the quarter-wavelength resonant frequency of the will show a slightly higher (by a few percent) electrical height for a given physical length. When a capacitive top hat or a Yagi or quad beam is mounted on top of the vertical, the electrical height will be increased considerably.

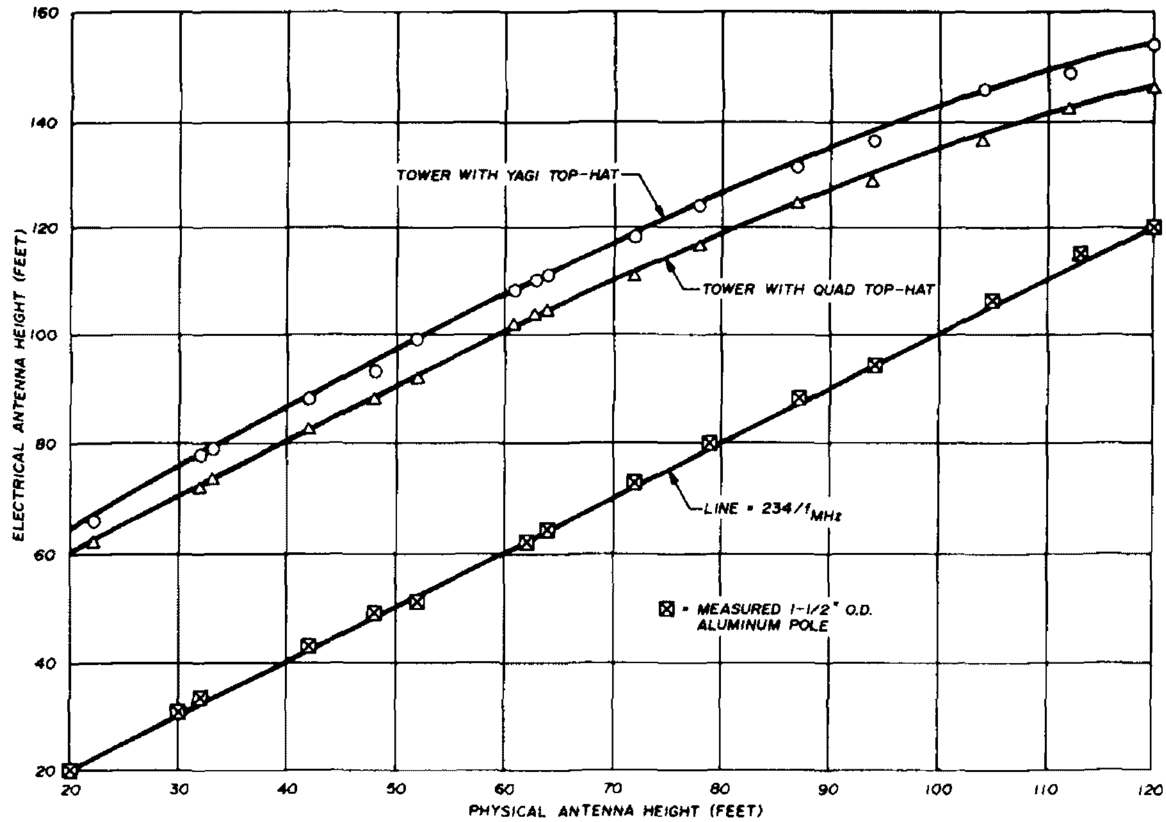


fig. 11. Physical vertical antenna height versus electrical height. Capacitive top-hat loading, as provided by a Yagi or quad, increases electrical height considerably.

vertical, the impedance will decrease to a minimum, indicating a null in the horizontal scope trace. When the frequency marker is accurately centered in the null as shown in **fig. 8** the quarter-wave resonant frequency is indicated.

The height of a quarter-wavelength vertical antenna operating against a good ground is given by

$$\text{height (feet)} = \frac{234}{f_{MHz}}$$

$$\text{height (meters)} = \frac{71.6}{f_{MHz}}$$

Vertical antennas made of small diameter materials will have a 1:1 relationship of physical length to electrical height when compared to this formula, but thick vertical elements, such as a metal pole one-inch (2.54 cm) or more in diameter

The curves plotted in **fig. 11** show that a 1½-inch (3.8-cm) diameter metal pole is electrically slightly longer than a thin wire of the same length. This chart also shows the lengthening effect that Yagi and quad beams have on the electrical height of towers.

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ham radio

three-band vertical DX antenna

Here's the result of some scrap materials and amateur ingenuity — a low-angle radiator for 40, 20 and 15 meters

Frank Lallak, W9BQE, 4061 North Drive, Fort Wayne, Indiana 46805

DX can be worked without a beam antenna. I've used a vertical antenna for the past seventeen years. Although I don't operate as many hours as most DXers, nevertheless I've worked my share of DX stations. The antenna described here is the result of some experimentation over the years. It will give a good account of itself.

Years ago I purchased parts for a 20-meter Yagi. Before I obtained the appropriate tower, TV became popular, and my neighbors decided they would rather view sub-fringe TV stations than the black-and-white interference bars caused by my rig. Scratch one Yagi antenna.

early attempts

My first vertical consisted of a 34-foot (10.4 meter) length of 3-inch (7.6-cm) diameter aluminum tubing, which was made from the boom for my Yagi. This arrangement blew down, but the only damage was a slight bend to the tubing. The next attempt was a supporting member for the tubing consisting of a 4 x 4-inch (10.2 x 10.2-cm) post, but unfortunately the post had a large knot at

ground level. The 4 x 4 was replaced with two 2 x 6-inch (5.1 x 15.2-cm) wood members bolted together. I attached the aluminum tubing to the wooden mast and added a 13-foot (4.0-meter) length of

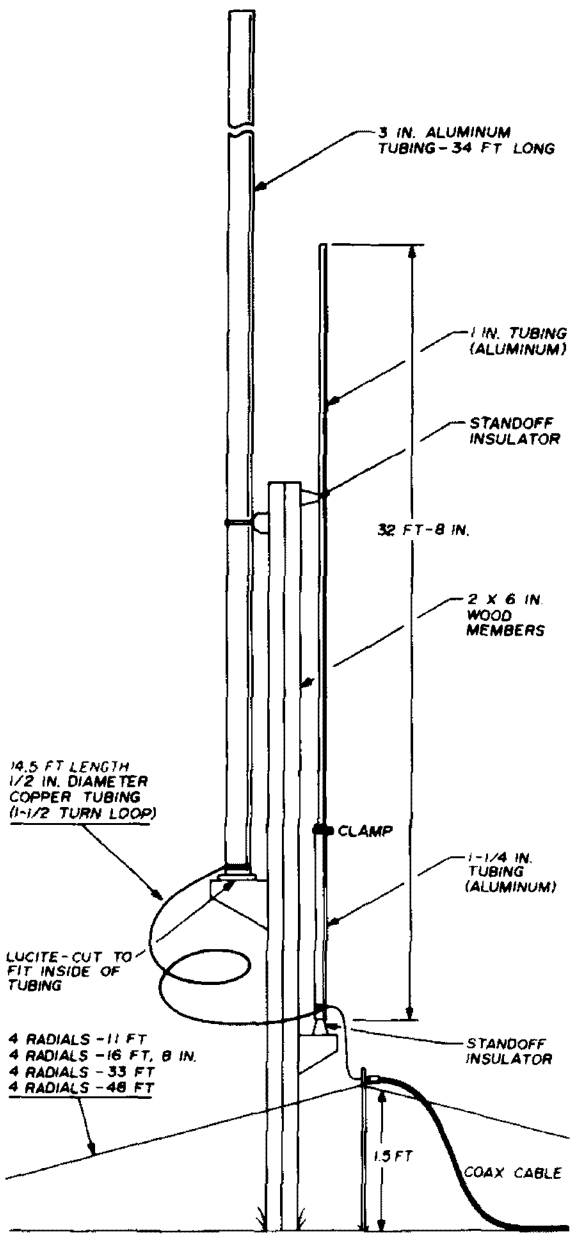

fig. 1. The three-band DX vertical. Additional radials will improve performance; the more radials, the better.

½-inch (13-mm) diameter copper tubing, formed into a 1½-turn loop at the bottom end, which was about six feet (1.8 meters) above ground. Two 32-foot (9.8 meter) radials completed the system. The copper tubing extended the antenna length to 3/4 wavelength on 20 meters, which produces a low vertical radiation angle.

This antenna worked fine on 20. To work the 40-meter band, I made a clamp and a double coax fitting, which I fastened at the 32-foot point on the 3-inch (7.6 cm) aluminum tubing. To change bands, I simply changed the coax cable from the 20-meter to the 40-meter spot on the tubing and connected the radials to the other side of the coax fitting. This arrangement produced a 1:1 swr on 40 meters; however, I wasn't able to work anything on 20 meters until I changed the coax to the 20-meter point. I used this two-band combination for several years but finally became tired of changing the coax on cold winter nights.

final design

The two-element vertical antenna I am now using is shown in **fig. 1**. The antenna on the right-hand side of the wooden mast consists of two pieces of tubing that were originally intended for elements on a Yagi antenna. The top section, of 1-inch 25.4 mm) aluminum tubing, telescopes into the bottom section, which is 1¼-inch (37.5 mm) tubing. Both sections are held together with a clamp. The bottom portion of the 20-meter antenna, which is the 1½-turn loop of copper tubing, is connected to the bottom end of the tubing on the other side of the mast. Radial wires, as shown in the figure, should be used for best results.

performance

The swr on 40 meters is 1.1:1. On 20, it is 1.1:1, and on 15 it is 1.25:1. I tried the antenna on 80 meters this last winter, but the swr was extremely high on the low end of the band. On the phone portion of 80, the swr is 3:1. I have had good reports, coast-to-coast, on 80 but haven't worked any DX. On the higher-frequency bands, the antenna has performed well. I've used the antenna for two years and am pleased with the results.

ham radio

160-meter loop
for receiving

Construction of a simple 160-meter loop antenna for receiving that reduces noise and strong-signal interference

Charles G. Bird, K6HTM, 875 Lindo Lane, Chico, California 95926

Reception on 160 meters can be greatly improved by using a loop antenna. Its primary virtue, reducing noise pick-up, is due to its small relative size and directional characteristics when properly located. A loop antenna is easy to build and doesn't have to be nearly as complicated as some loops that have been described.[1]

construction

A two-turn loop, 23 inches (58.4 cm) in diameter was fashioned out of a length of semi-flexible aluminum-sheathed coaxial cable. Braided cable would have been fine. The vinyl jacket acts as a dielectric, and the two turns are securely taped, the ends being brought close together and bridged with sufficient capacitance to resonate at 1.835 MHz. The center conductor is ignored. A voltage divider across the ends permits coupling to a low-impedance transmission line.

The 23-inch (58.4 cm) dimension resulted from a random length of cable and was used as a starting point. To obtain a rough idea of the inductance of this loop I used the following formula:

$$L = \frac{R^2 N^2}{9R + 10S}$$

Where L is the inductance of the coil in microhenries, R is the radius of the coil in inches, S is the length of the coil in inches and N is the number of turns. For the two-turn, 23-inch loop I built out of coax,

$$L = \frac{(11.5)^2 (2)^2}{(9 \times 11.5) + (10.1)} = \frac{529}{113.6} = 4.66 \mu H$$

Although this value may not be particularly accurate, it is quite helpful in determining an experimental value of capacitance at C1 to resonate the loop to 1.835 MHz. From the resonance table in the *Radio Amateur's Handbook*[2] this works out to be about 1600 pF.

By adjusting capacitor C1 it is possible to peak the response over a narrow portion of the 160-meter band. Tuning your receiver across the noise hump provides the surest indication of the resonance point. In **fig. 1** the resonating capacitance includes the parallel capacitance of the input voltage divider. Using the equivalent parallel capacitance as a basis (approximately 1424 pF), the actual inductance of the loop, at 5.28 μH, is slightly higher than that given by the formula.

At 1.835 MHz a 1735-pF capacitor exhibits 50-ohms reactance. I used the closest value I had available, 1670 ohms. A 100-pF capacitor provided satisfactory coupling for the 50-ohm transmission line.

performance

Compared to receiving on a half-wave dipole, the loop is like turning on a light. Both the noise and strong signals are reduced by 20-dB or more, but more important, unreadable signals become clearly readable. This result was enough. The directional characteristics which were noted at the test bench disappeared when

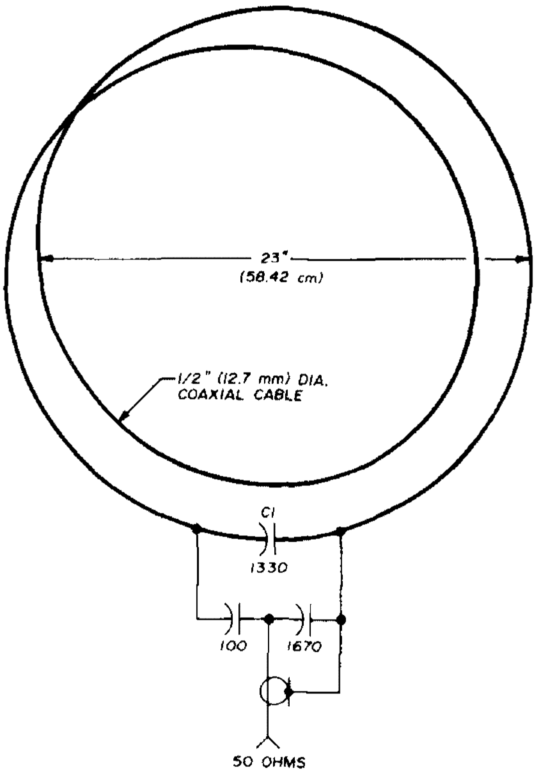

fig. 1. Simple 160-meter loop antenna for receiving.

the loop was suspended from the boom of a quad at eighty feet (24.4 meters). Those operators who are bothered by noise from a specific location or QRM from a certain direction would do better to mount the loop away from any reflecting objects and rotate it accordingly.

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ham radio

the truth about 5/8-wavelength vertical antennas

This practical discussion of 5/8-wavelength vertical antennas adds perspective to the varied gain claims seen in the amateur magazines

Paul Meyer, KØDOK, 9921 Crestwick Drive, St. Louis, Missouri 63128

The recent growth in popularity of the 5/8-wavelength vertical radiator, particularly for use at vhf, has given rise to many varied claims as to its performance. Several local amateurs have been unable to detect improved performance after changing from quarter-wave to 5/8-wavelength antennas. Because of my vocation as an antenna engineer, I was prevailed upon to examine the problem. A theoretical analysis was considered and several actual measurements were performed.

theory

A monopole antenna mounted on a ground plane may be considered equivalent to a free-space vertical dipole having each dipole arm equal to the length of the monopole. The lower arm of the dipole corresponds to the image of the monopole reflected in the ground plane. Equivalent radiation patterns may be expected as shown in **fig. 1**. A ground plane of infinite size is assumed.

Actually, with any finite ground plane size and where the field strength is measured at a point some distance beyond the edge of the ground plane, the pattern is altered considerably by reflection from the discontinuity at the edge of the ground plane. The resulting pattern has multiple lobes caused by phase interference of the direct and edge reflected waves. **Fig. 2** shows the results of an actual measurement made with a large but finitely-sized ground plane. Notice that the signal maximum is at an angle considerably above the horizon where it is not useful for ground-wave communications.

If an infinite ground plane is assumed and the monopole *is analyzed as its*

equivalent dipole, we must formulate a mathematical expression for the field strength as a function of angle from the antenna. This angle, commonly called θ, is measured from the zenith (straight up) and equals 90 degrees at the horizon. The field strength at any θ angle may be calculated by summing the radiation from an infinite number of short current elements spaced evenly along the length of the dipole. This type of summation is performed mathematically by a calculus technique known as integration.

Phase must be considered in this summation since the radiation from some current elements may cancel that from others by being out of phase. The amplitude of the currents in the short elements are not equal but each is a function of its position along the antenna. A very close approximation is to assume the current to be sinusoidal starting from zero at the top end because the end is an open circuit and no current can exist there.[1]

Assuming a sinusoidal current distribution along the dipole and performing the mathematical integration yields the formula for field strength of a dipole:[2]

$$E = \frac{60}{r} \times I_o \times \frac{\cos(\pi L \cos\theta) - \cos \pi L}{\sin\theta}$$

Where

E = field strength in volts per meter

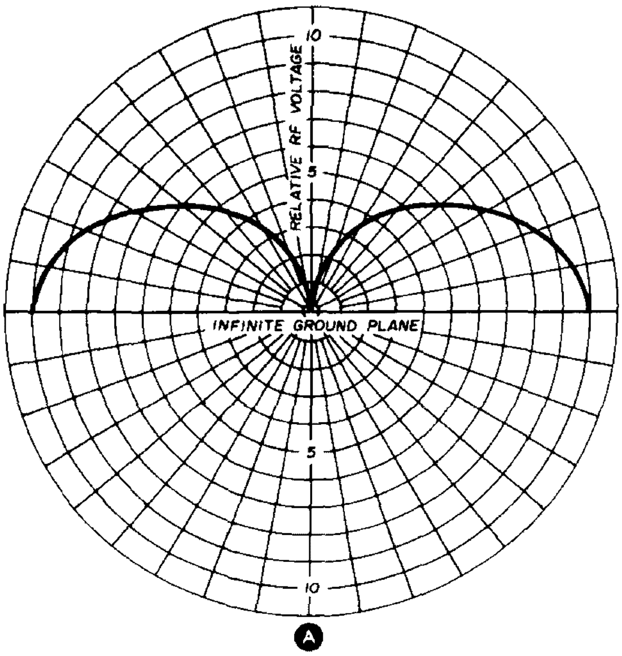
I_o = current maximum in rms amperes

L = overall length expressed in wavelengths

r = distance from the antenna in meters

The mathematically inclined reader will recognize that for a given current maximum, wavelength and distance this formula yields a maximum horizon (θ=90°) field strength when the dipole is one wavelength long. If you intend to use an impedance-matching network to match any arbitrary length dipole to the transmission line, and if it is assumed that the network is lossless, the transmitter will be able to deliver the same power to the antenna regardless of its length. However, when a constant radiated power is maintained, the antenna current (also the unmatched impedance) depends upon the length of the antenna. The optimum dipole length for maximum horizon field strength will, therefore, not be one wavelength but will be somewhat longer.

Since the formula is in terms of a constant rms current maximum, the problem of determining optimum length for maximum gain becomes complicated.



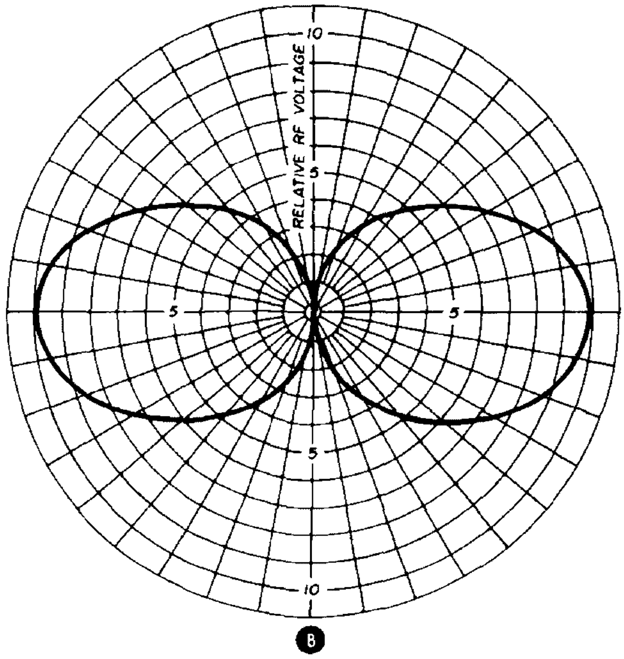


fig. 1. Theoretical radiation pattern of a quarter-wave vertical monopole on an infinite ground plane is shown in (A). Theoretical pattern of a half-wave dipole in free space is shown in (B).

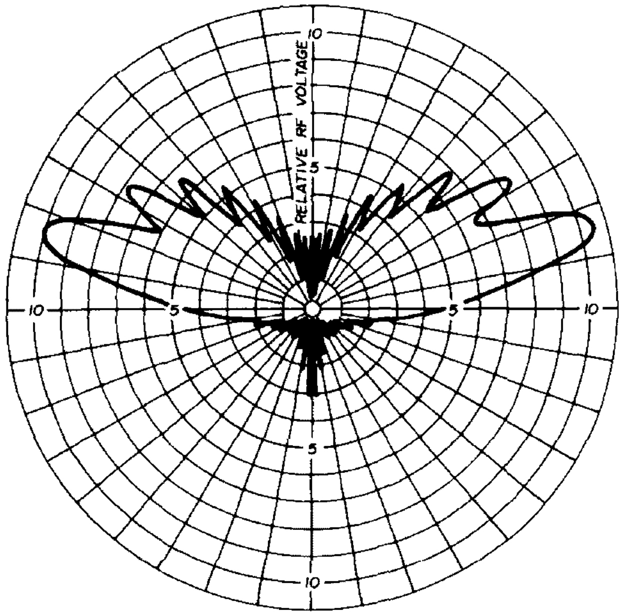


fig. 2. Radiation pattern of a quarter-wave vertical mounted in the center of a ground plane 16 wavelengths in diameter. (Antenna range test at 8 GHz, quarter-wave stub = 0.37", ground plane, 24.4" diameter.)

One approach to the problem is to construct the three-dimensional antenna radiation pattern from the formula and then calculate the total power contained within the radiated field. You may then adjust *current* I_o as necessary to hold the power constant as the antenna length is varied. The horizon field strength may thus be calculated under these conditions for various lengths until the optimum length is determined.

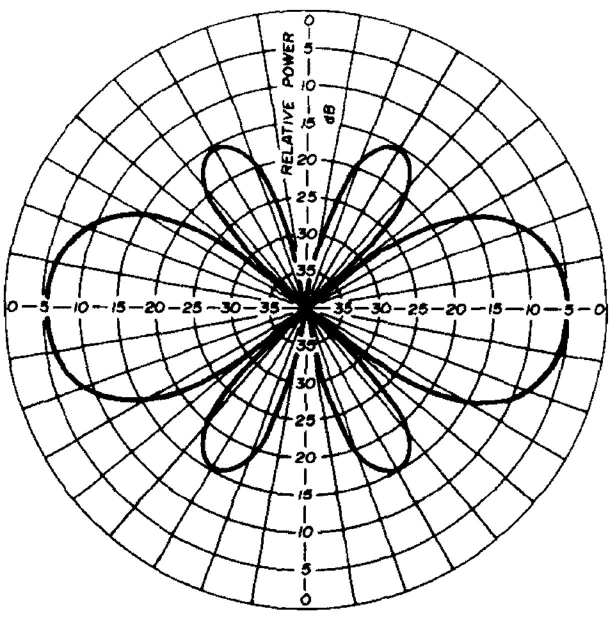


fig. 3. Theoretical radiation pattern of a 1.27-wavelength dipole in free space.

In order to do this, you must have a method to calculate the power contained in the radiated field. Such a method exists if you can use the calculus technique of integration.[3] Considering that the entire sphere of all directions from the antenna can be made up of an infinite number of small segments of solid angle and that the power contained in each segment is proportional to the square of the field strength (voltage) in that particular direction, a mathematical summation (integration) may be performed to calculate the total power.

The result of this calculation may be obtained as an isotropic value. The isotropic is the field strength which would result in all directions from the antenna if the radiated power were spread evenly over the entire sphere about the antenna rather than being concentrated in certain directions as is the actual case.

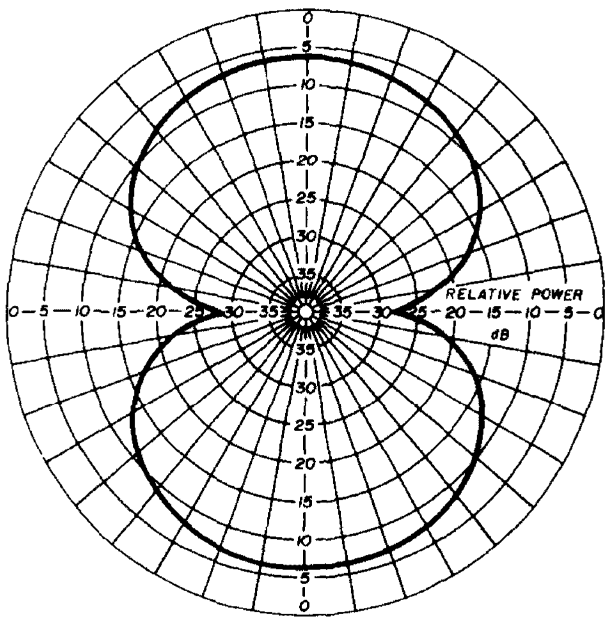


fig. 5. Radiation pattern of the balun-fed and matched half-wave dipole shown in fig. 4A.

Details of the calculus procedures are beyond the scope of this article. However, I have calculated the results shown in table 1, assuming the same current regardless of the dipole length. The results show that the half-wave dipole itself has directional properties and exhibits *2.15-dB gain over an isotropic* (non-directional) radiator. This is an important fact since the gain of any antenna may be

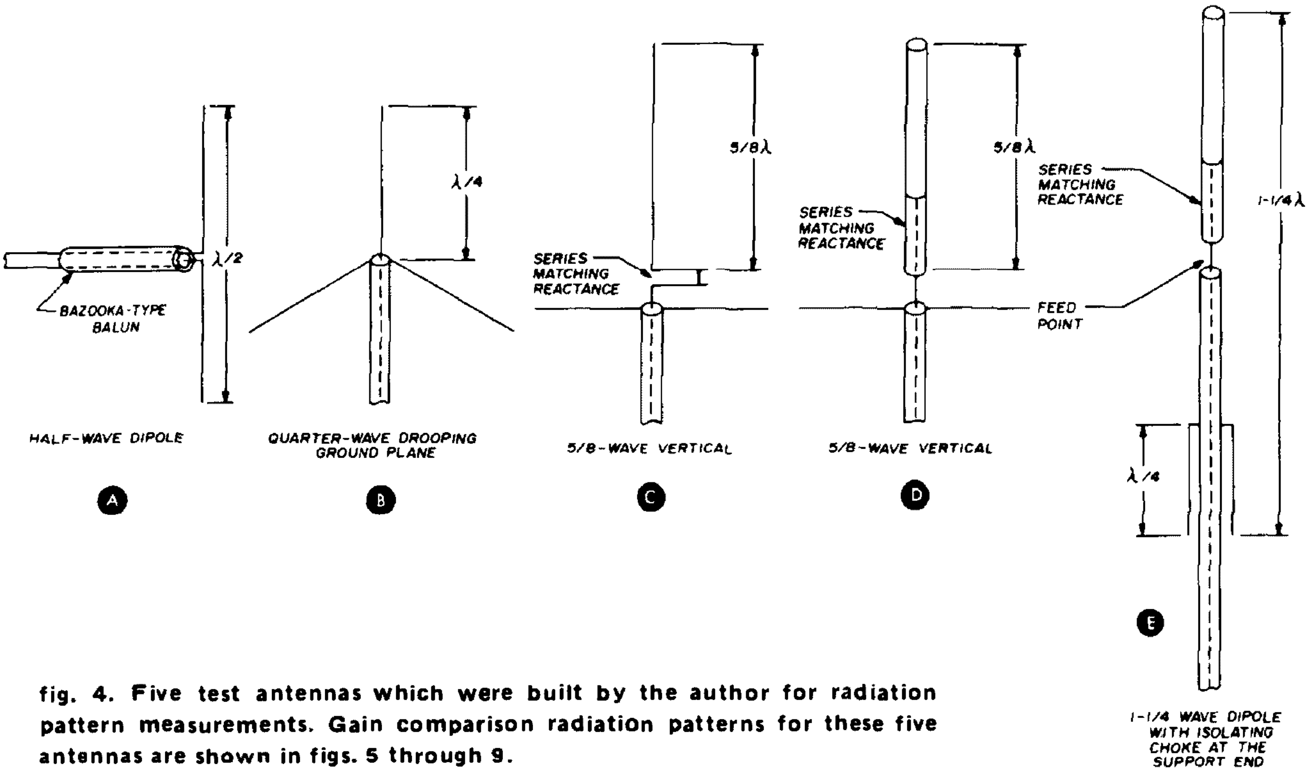


fig. 4. Five test antennas which were built by the author for radiation pattern measurements. Gain comparison radiation patterns for these five antennas are shown in figs. 5 through 9.

expressed in dB above a dipole, or isotropic, or some other standard. Using the isotropic as reference will, of course, make the gain figure 2.15-dB higher than if a dipole is used, and larger numbers look better in the advertising. When you read a manufacturers claim be sure that the gain reference is specified, otherwise the claim is meaningless.

The results also show that the dipole reaches its maximum gain when its length is 1.270 wavelengths, and that the gain decreases as it is further lengthened. That maximum of 5.18 dBi (dB over an isotropic) is equal to 3.03 dBd (dB over a dipole).

Thus, we have verified the basis upon which the 1-1/4-wave dipole or equivalent 5/8-wavelength monopole radiator is commonly considered to be a 3-dB gain antenna. Bear in mind that this is the theoretical maximum gain for the ideal situation of a matched dipole in free space or an equivalent matched monopole on an infinite ground plane. If the ground plane is less than infinite, the gain will be reduced from this maximum value. A plot of the theoretical radiation from a 1.27-wavelength dipole appears in fig. 3.

actual measurements

In an attempt to determine just how much gain reduction results from using ground planes of practical dimension, I built a number of antennas and carefully measured their radiation patterns. The antennas were measured on a test range at 1000 MHz where signal reflections from the ground and surroundings could be controlled to at least 40-dB below the incident signal. The antennas were rotated while signal strength was auto-

table 1. Field strength and gain of various length horizontal dipoles.

Dipole (wavelengths)	Field (volts/meter)	Isotropic (volts/meter)	Gain Ratio (over isotropic)	Ratio (dB)
0.500	1.00	0.78065	1.2815	2.150
1.265	1.6730	0.92165	1.8152	5.179
1.270	1.6613	0.91511	1.8154	5.180
1.275	1.6495	0.90878	1.8150	5.178

matically plotted on a polar graph recorder.

Five antennas were constructed and their radiation patterns measured. The antennas were a half-wave dipole (fig. 4A), a quarter-wave vertical with a drooping four rod ground plane, (fig. 4B), a 5/8-wavelength vertical above a plane of four 1/4-wavelength rods and using a shorted section of open-wire line in series with the feed to cancel reactance (fig. 4C), a similar 5/8-wavelength vertical except with a shorted coaxial section for

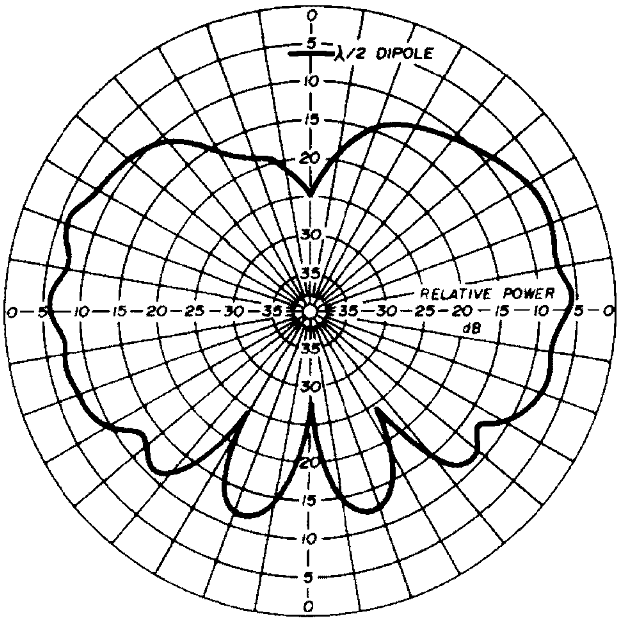


fig. 6. Radiation pattern of the quarter-wave drooping ground plane antenna shown in fig. 4B.

the series matching reactance (fig. 4D), and a 1-1/4-wavelength vertical dipole with a coaxial series matching reactance and end support isolated by a quarter wave choke (fig. 4E). All antennas were matched to less than 1.2:1 vswr so that no mismatch losses had to be considered when making gain measurements by substituting one antenna for another. The radiation patterns are shown in figs. 5 through 9.

As expected, the half-wave dipole and the quarter-wave drooping ground plane exhibited nearly the same gain. Although the 5/8-wavelength verticals produced somewhat smoother patterns, they failed to yield significant gain over the dipole or quarter-wave vertical. The two schemes

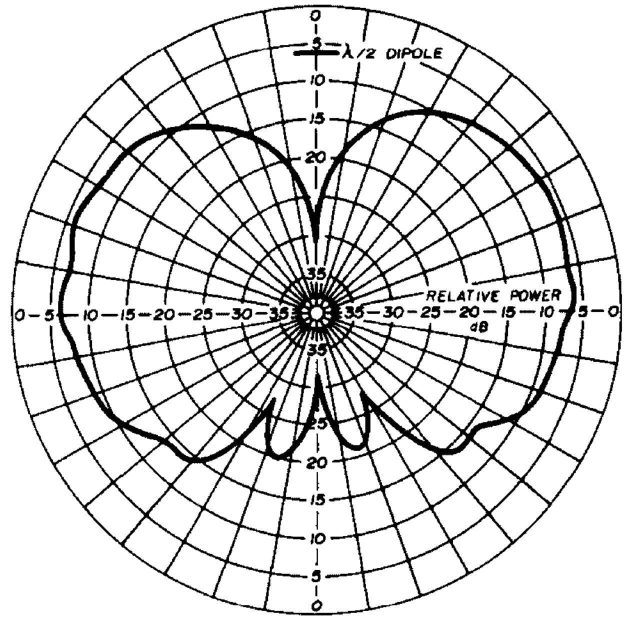


fig. 7. Radiation pattern of the 5/8-wave-length vertical on a 4-rod quarter-wave ground plane with a hairpin matching loop shown in fig. 4C.

for matching the 5/8-wavelength rods appear to be equivalent. The 1-1/4-wavelength vertical dipole antenna supplied nearly the theoretical 3-dB gain and exhibited a pattern shape similar to that predicted from the formula.

My only conclusion from this testing is that the usual image plane analysis of vertical monopoles is valid only for infinite ground planes and is greatly in error

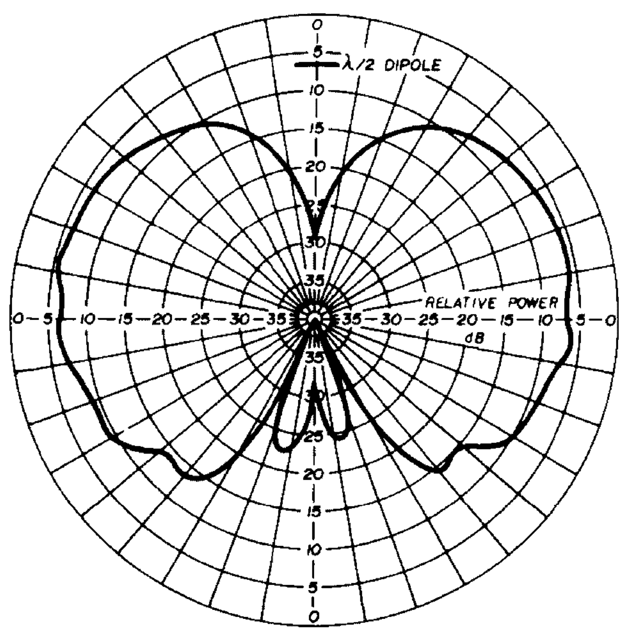


fig. 8. Radiation pattern of the 5/8-wavelength vertical on a 4-rod quarter-wave ground plane with a series coaxial matching reactance shown in fig. 4D.

for very small ground planes. As you can see from **fig. 1**, even a ground plane many wavelengths in size is a poor approximation of the infinite image plane. Therefore, it is not surprising that despite many gain claims, the 5/8-wavelength vertical ground plane frequently disappoints those expecting performance exceeding that of a quarter-wave ground plane. The mobile antenna situation is somewhat better due to the larger ground plane that a rooftop provides, but it is doubtful that even this advantage can add up to the 3-dB gain theoretically achieved on an infinite ground plane.

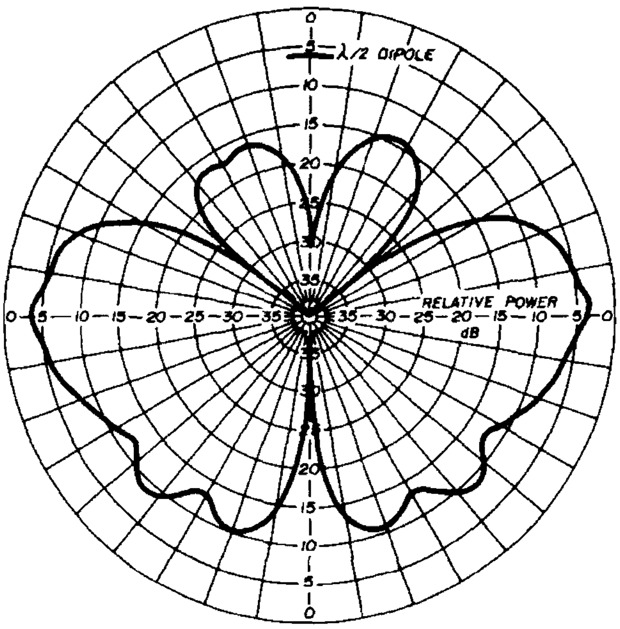


fig. 9. Radiation pattern of the matched 1-1/4-wavelength sleeve dipole with a decoupling choke shown in fig. 4E.

It is noted that the FCC is now requesting certified gain and pattern measurements from the manufacturers of antennas intended for use in amateur repeater stations.[4] The results of those measurements should indeed be very interesting.

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1. J. Kraus, *Antennas,* McGraw-Hill, New York, 1950, chapters 5.5 and 9.8
2. H. Jasik, *Antenna Engineering Handbook,* McGraw-Hill, New York, 1961, chapter 2.1.
3. J. Hollis, *Microwave Antenna Measurements,* Scientific-Atlanta, Inc., 1970, chapter 7.
4. "FM Repeater News," *QST,* April, 1973, page 94.

ham radio

five-to-one transmatch

Harry R. Hyder, W7IV, 9842 N. 57th Street, Scottsdale, Arizona 85253

Transmatches always seem more complicated than they have to be, particularly if you recall that *any transmission* line impedance, real or complex, at *any* vswr, can be made to look like a 50-ohm resistive load by the use of only one inductance and one capacitance in the familiar L-network shown in **fig. 1.**

What complicates the design of most transmatches is the desire to match almost any impedance with reasonable and practical values of L and C. Examples of extreme mismatch problems are matching a "random length" long-wire antenna, using a center-fed dipole on half or twice its resonant frequency or using an antenna fed with open-wire transmission line on several bands.

These are extreme cases. More common problems are the need to match a beam antenna that has been carefully pruned for the phone band in the CW portion, or the need to use a single coax-fed antenna over the entire 3.5- to 4.0-MHz band. The mismatch in these cases is usually more than the transmitter's output network can handle, but in general the vswr is not more than 5:1.

If you were to design a transmatch to handle a maximum vswr of 5:1 on a 50-ohm line, it could be made quite simple, and would not require more than one inductance and one capacitance, both of quite practical size.

As amateurs should all know by now, vswr for *practical* lines on bands below 30 MHz is independent of line length. The resistive and reactive components of the mismatch, however, are a strong function of line length. For a vswr of 5:1 on a 50-ohm line of any length, the resistive component will be between 10 and 250 ohms and the reactive component will be between plus and minus 120 ohms. A transmatch that will match these resistances and cancel these reactances on the 3.5- to 30-MHz bands is diagrammed in **fig. 2.**

construction

The rotary inductor could be a Johnson 229-202 or equivalent, having an inductance of 18 μH. The capacitor is a surplus 5-gang unit used in WWII direction-finding receivers. This capacitor has 420 pF per section (2100 pF total with all in parallel). It should be used with a switch (S2), so that only one or all five sections can be used. The minimum capacitance with all five sections in parallel approaches 100 pF, and less than this may be needed. This capacitor is available from Barry Electronics and Fair Radio Sales, and possibly other surplus dealers.

A word about component ratings. The capacitor is a high-grade receiving unit, but for any *legal* power rig, the voltage across it is nearly 1000 volts peak. I have tested my own unit to 1500 volts dc without any arc over. However, you may get a higher voltage than this is you are in the habit of tuning up at full power. If the capacitor does break down, it will impress on you forcibly that this is very bad practice and should be discontinued.

The roller inductor is rated at 5 amperes. Under some conditions the current may be higher than this, but not dangerously so. Here, the problem is heat; excessive heating could loosen the wire on the ceramic form or melt a solder joint. Except for high-power sstv or RTTY, average current will always be less than 5 amperes.

The switches should be heavy-duty rf types. The switches from surplus BC-375

tuning units are good, as are the excellent rf switches made by the James Millen Company.

tune up

The best way to tune up this, or any other transmatch, is with a low-power 50-ohm bridge and a signal source such as the Omega-T Noise Bridge, or the simple resistive bridge described in most editions of the *ARRL Handbook,* using a grid-dip meter as the source.

If all you have is a vswr meter and you don't want to invest in any more test equipment, you should use no more power than is necessary to obtain a reliable vswr reading. Tuning up on full power is ungentlemanly, illegal and hard on components. Of course, when the proper settings have been found for all your mismatch conditions, they should be recorded for future operation under those conditions.

Connect your vswr bridge between the power source and the transmatch. Since I must presume that all you know is your vswr and have no idea of what resistance and reactance it represents, you can flip a coin as to the initial position of S1. Switch S2 should place all sections of the variable capacitor in parallel.

Start with an inductance of zero, then rotate the capacitor through its range. Make small increases in inductance, tuning the capacitor each time until a vswr of 1:1 is obtained. If none is found, throw S1 to its other position and repeat the process. This is far less tedious than it sounds. The tuning is very broad, because the Q never exceeds 2.25. If you think that the Ls and Cs you get for a match don't seem to be proper for the band in use, don't worry. This could be a point where the resistive component is close to 50 ohms, but where the reactance is as high as plus or minus 90 ohms—all perfectly normal.

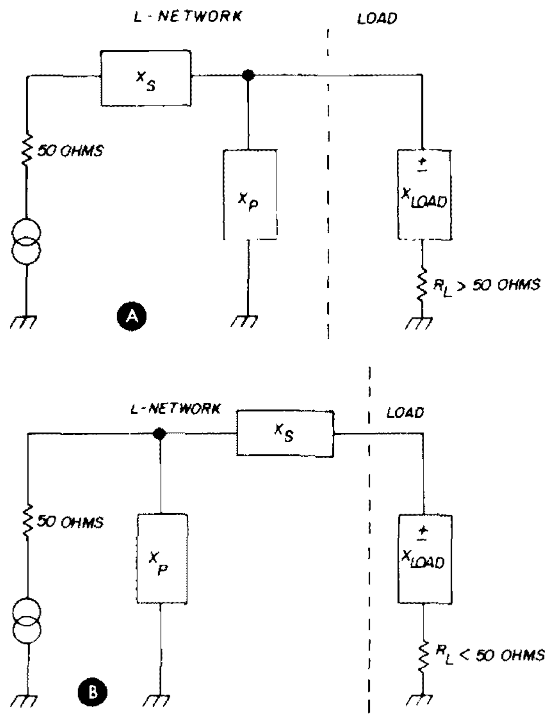

fig. 1. Basic L-network impedance-matching circuits. Circuit in (A) matches loads with resistive component greater than 50 ohms. L-network in (B) matches loads with resistive component less than 50 ohms.

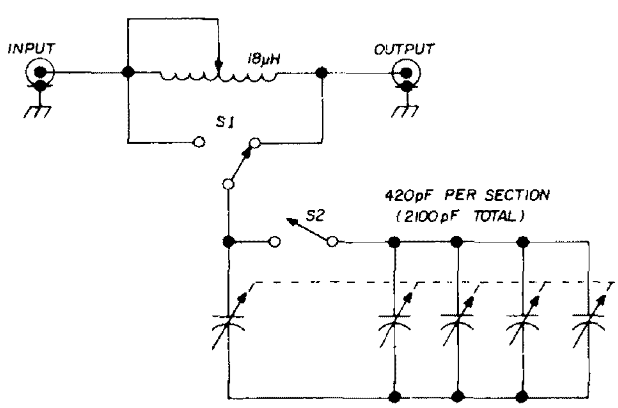

fig. 2. Circuit for the five-to-one transmatch. The rotary inductor is an E.F. Johnson 229-202 or equivalent.

summary

The five-to-one transmatch will actually handle some, but not all, mismatches greatly in excess of 5:1 on the 3.5-MHz band. On the 7-MHz band, the range is far higher, and on higher frequencies it is almost unlimited. However, the components, particularly the variable capacitor, can easily fail at some standing-wave ratios higher than 5:1 when used with a high-power rig. This should be taken into account.

In any event, if your mismatch requirements are moderate, you can easily build this transmatch at a fraction of the cost of a wider-range unit.

ham radio

vertical radiation

patterns of horizontal antennas

How antenna height affects the vertical radiation angle of horizontally polarized antennas

H. Glenn Bogel, WA9RQY, 4111 Beaverbrook Drive, Fort Wayne, Indiana 46805

There are many sources of information describing the vertical radiation angles of horizontal antennas (the angle above the horizon of the axis of the main lobe of radiation), but most of these references tend to go so deeply into theory that the reader is unable to answer the basic question, "How can I determine the radiation angle of *my* antenna?" The graphs presented in this article will go a long way toward answering that question.

Height above ground, not antenna type, determines the radiation angle of horizontally polarized antennas. Although a parasitic antenna, such as a Yagi, will cause more power to be radiated at a low angle than a dipole, in both cases the vertical radiation angle is the same, assuming both antennas are at the same height. The effects of ground reflection are probably most easily understood by studying the *image antenna* approach found in the *ARRL Antenna Book*[1] and other publications.[2]

Basically, if the antenna were to be suspended in free space, the main lobe of radiation would be directly in line with the aperture of the antenna. When the antenna is located near the earth, however, a conflict occurs between the direct wave from the antenna and the wave that is

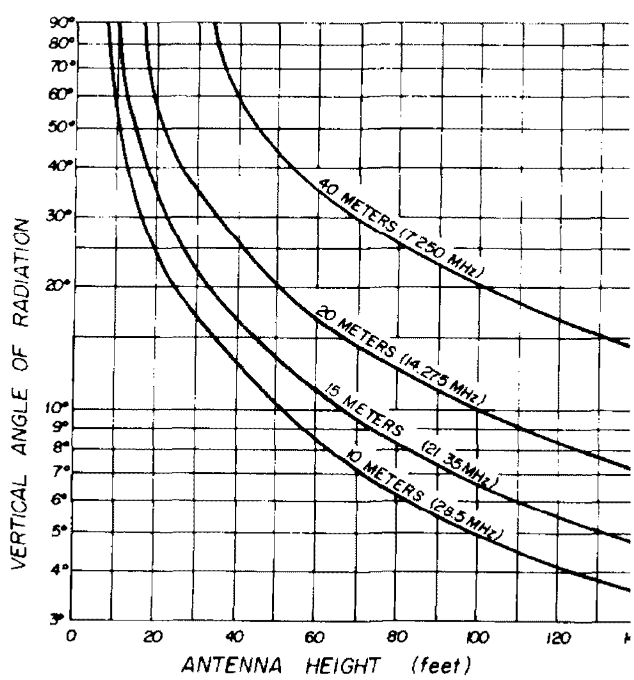

fig. 1. Graph of major vertical radiation lobes for various horizontal antenna heights up to 140 feet.

reflected from ground. The phase difference between these two waves results in cancellation and reinforcement at various angles above the horizontal. Where reinforcement occurs lobes are present, and at points of cancellation nulls appear in the vertical radiation patterns. Varying

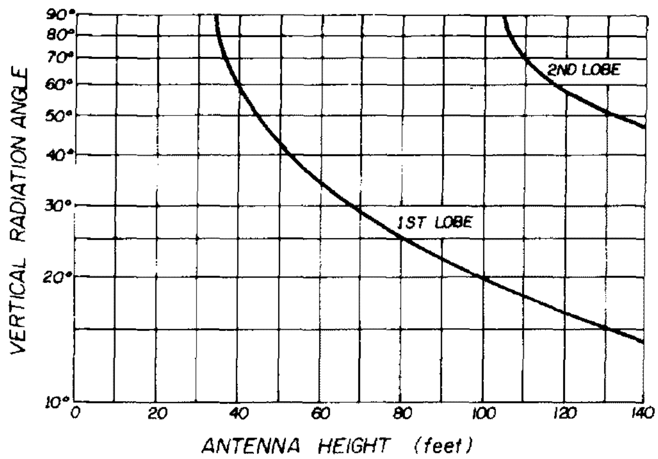

fig. 2. Vertical radiation angle of first and second lobes of a 40-meter (7.250 MHz) horizontal antennas at heights up to 140 feet.

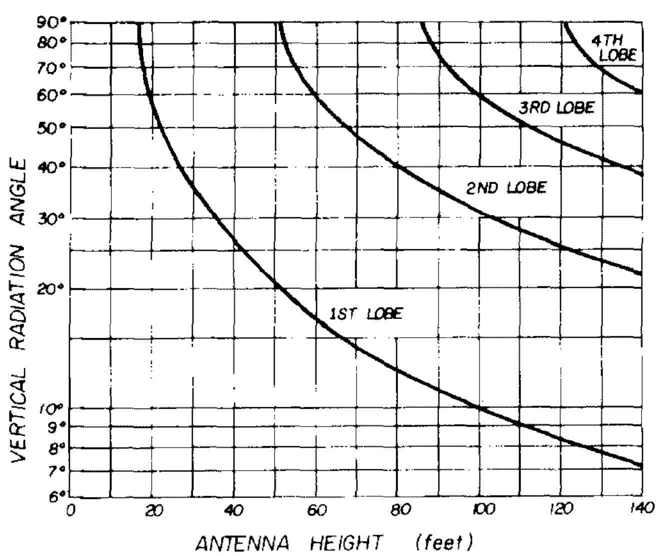

fig. 3. Vertical radiation angle of pattern lobes for 20-meter (14.275 MHz) horizontal antennas at heights up to 140 feet.

the height of the antenna above the reflecting ground changes the vertical angles of cancellation and reflection patterns.

My interest in DXing, contest operating, rag chewing and antennas led me to a very practical magazine article on the subject of vertical radiation patterns.[3] This, in turn, resulted in many days of calculations and graph making, producing the graphs presented here which show angle of radiation vs antenna height above ground for 10, 15, 20 and 40 meters.

The following method was used to plot the graphs. This approach is based on the fact that the incident and reflected waves, when analyzed vectorially, are in phase at 90°, 270°, 450°, 630°, 810°, etc.

1. Calculate the wavelength (in feet) of the frequency being considered:

$$\lambda = \text{wavelength (feet)} = \frac{984}{f_{MHz}}$$

This represents one wavelength or 360 electrical degrees.

2. Determine the antenna height in electrical degrees:

$$h_A = \text{antenna height (degrees)} = 360\frac{h_1}{\lambda}$$

where h_1 is the height of the antenna and λ is one wavelength, both in feet.

3. Compute the vertical angle of radiation for each lobe

$$\sin\alpha = \frac{90}{h_A}\ \frac{270}{h_A}\ \frac{450}{h_A}\ \frac{630}{h_A}\ \frac{810}{h_A}\ \text{etc.}$$

where α is the vertical angle of each lobe.

4. From a table of natural trigonometric functions, find the vertical radiation angle, α, for each value of $\sin\alpha$ up to 90°. Since 90° is straight up ($\sin\alpha$ = 1.00), higher values are not valid.

For example, assume you have a horizontal 20-meter antenna installed on top of a 70-foot tower. What is the vertical angle of radiation at an operating frequency of 14.275 MHz?

1. One wavelength at 14.275 MHz is 68.9 feet:

$$\lambda = \frac{984}{14.275\ \text{MHz}} = 68.9\ \text{feet}$$

2. A 70-foot tower represents 365.6° at 14.275 MHz:

$$h_A = (360°)\frac{70}{68.9} = 365.6°$$

3. To find the vertical angle of the first

(primary) lobe,

$$\sin \alpha = \frac{90^\circ}{365.6^\circ} = 0.246,$$

Consulting a table of natural sine functions, the vertical angle is approximately 14.25° (see fig. 3). The second lobe is at approximately 47.5°.

$$\sin \alpha = \frac{270^\circ}{365.6^\circ} = 0.739, \alpha = 47.5^\circ$$

Therefore, the first (major) lobe occurs at 14.25° above the horizontal while the vertical angle of radiation for the second lobe is 47.5°.

Fig. 1 is a composite graph of the first, or major lobes, on 10, 15, 20 and 40 meters for antenna heights from 8½ to 140 feet. The curves were all calculated for the center of the American phone bands except for 10 meters. The low end of the American portion of 10 meters was chosen because of the concentrated activity in this area of the band. The angle of radiation will not vary appreciably from one end to the other on any of the bands.

Figs. 2 through 5 show all of the lobes present on each band for any given height. Again, the lines are the lobes of vertical radiation and the null points are approximately midway between lobes. Never pick an antenna height which will present any lobe at 90° – this results in wasted radiated power. One interesting observation I made is that 70 feet is the only antenna height on the graphs that has a null point at 90° on each band!

It should be pointed out that these graphs were based on the assumption that electrical ground (the electrical plane from which the antenna waves are reflected) is at the physical surface of the ground. In actuality, electrical ground varies considerably from one location to another, and may be located from several inches to several feet below the surface.

These graphs, combined with a little operating experience, should be helpful in selecting the proper antenna height for DXing or short-haul propagation, or both, depending on your choice of radiation angle.

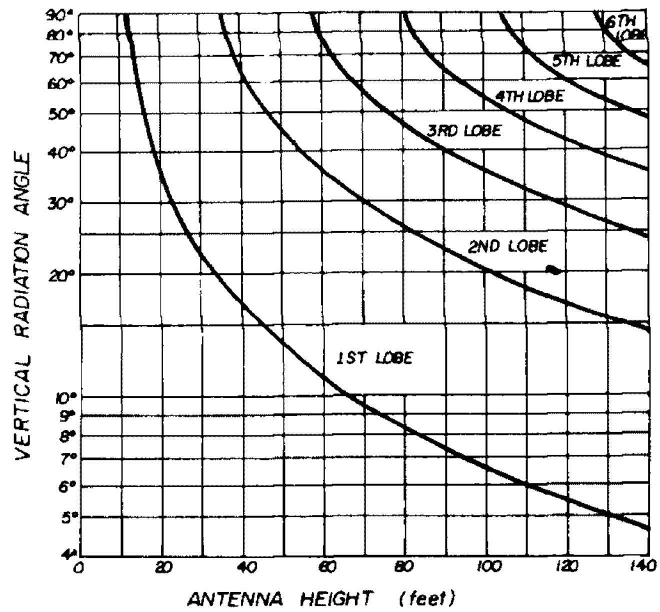

fig. 4. Vertical radiation angle of pattern lobes for 15-meter (21.250 MHz) horizontal antennas installed at heights up to 140 feet.

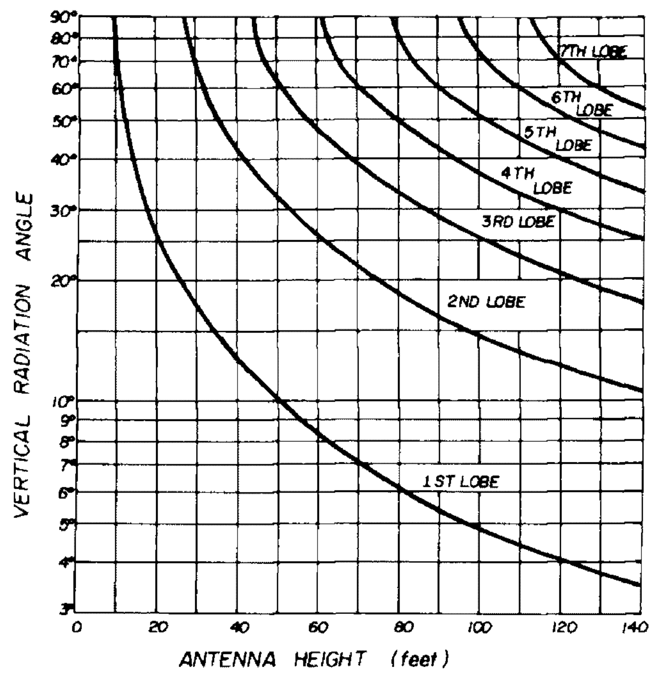

fig. 5. Vertical radiation angles of pattern lobes for 10-meter (28.5 MHz) horizontal antennas installed at heights up to 140 feet.

references

1. *ARRL Antenna Book*, 12th edition, American Radio Relay League, Newington, Connecticut, 1970, page 44.
2. William I. Orr, W6SAI, *Beam Antenna Handbook*, 4th edition, Radio Publications, Inc., Wilton, Connecticut, 1971, page 13.
3. Kenneth Schofield, W1RIL, "What's Your Angle," *CQ*, March, 1968, page 41.

ham radio

pi network design aid

Presenting a set of curves for choosing power amplifier components to provide optimum tank-circuit efficiency

Alf Wilson, W6NIF, 1068 Arden Drive, Encinitas, California 92024

The tank circuits of yesterday, with their beehive insulators and cumbersome copper tubing coils, have all but disappeared from modern transmitters. The main reason is, of course, that today's rigs are designed for use with coax transmission lines. The simplest and most efficient means of transforming the high plate-load impedance of the final amplifier tube to low-impedance coax is with a pi or pi-L network. The pi network is the most often used of the two circuits since it requires fewer parts.

Much design data on pi networks has appeared in the literature over the past 15 years and will not be repeated here. The purpose of this article is to present design data for those who wish to build a pi-network amplifier without wading through a multitude of table lookups and without having to convert component reactance values into equivalent capacitance and inductance.

ground rules

To keep things simple and still provide useful data for most amateur applications, the curves are based on the general case; i.e., a tank circuit Q of 12, tube load impedances between 1000 and 4000 ohms, and an output impedance of 50 ohms. The advantage of using these curves is that no interpolation is required for values that don't appear in published tables of such data. You can pick off

exact values of capacitance or inductance directly from the curves.

The curves are based on calculations for tube load impedances used in the Class C mode:

$$R_L = \frac{E_B}{2\,I_B} \qquad (1)$$

where

R_L = tube load impedance (ohms)
E_B = tube plate potential (volts)
I_B = tube plate current (amperes)

using the curves

As mentioned previously, the curves are based on an operating Q of 12, which is optimum in terms of tank-circuit efficiency and harmonic attenuation. Most rigs today cover the bands between 3.5 and 28 MHz with bandswitching coils.

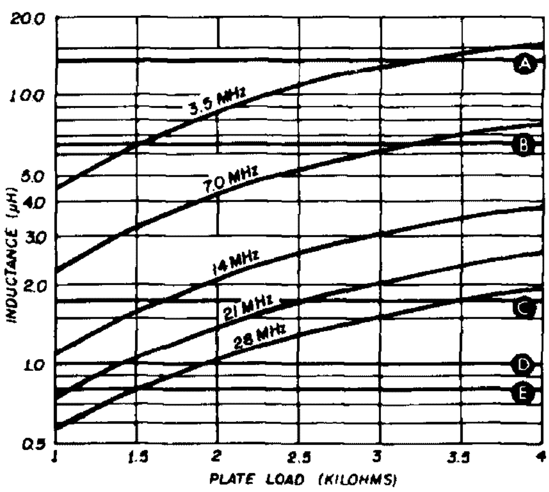

fig. 1. Curves of inductance as a function of tube plate-load impedance for pi networks. Loci of constant inductance for the five amateur hf bands are shown for the B&W 850A tapped inductor example discussed in the text (lines A, B, C, D and E).

One of the problems in such transmitter designs is finding components that will operate satisfactorily over the desired five-band frequency range while still maintaining reasonable circuit Q.

If the Q is to be held at or close to 12 in an amplifier that's switchable over the five high-frequency bands under different operating modes (CW and ssb), then it is difficult if not impossible to find off-the-shelf components for the pi network input capacitor, inductor and output capacitor. Several compromises have been reported for resolving this problem. An example is given in reference 1, which shows one way to deal with the designer's dilemma involving different plate-load impedances in a pi network tank for 1-kW CW and 2-kW pep ssb operation.

In this case, a multi-tapped coil is used, with two taps for each band, one for CW and one for ssb. Lacking a continuously variable inductor, this is one way to obtain the optimum inductance for different plate-load impedances — it requires a little work, but it's less expensive than buying a rotary inductor that will handle the power involved.

design example

An inductor can be designed to provide the correct inductance for most plate load impedances with the aid of the curves in fig. 1. Included in fig. 1 are loci of constant *inductance* (straight lines) for the popular B&W model 850A tapped bandswitching inductor. This inductor is extremely rugged and has been used by many amateurs in multiband amplifiers. However, as with many manufactured components, it is a compromise; the manufacturer tries to put out a product that will be useful for general applications, and to obtain optimum performance for a specific plate load impedance requires further work by the amateur.

For example, suppose you wish to design a five-band rf amplifier around the B&W 850A inductor. Assume that dc power input is to be 1 kW. If plate voltage is, say, 3000 volts and plate current is 0.3 ampere, the plate load impedance, from eq. 1, will be R_L = 1800 ohms. From fig. 1 inductance values for this load impedance would be:

band	L (μH)	B&W 850A L (μH)
3.5	7.6	13.5
7	3.8	6.5
14	1.88	1.75
21	1.25	1.0
28	0.92	0.8

Clearly, the taps on the 850A inductor are not located to provide the optimum inductance for the plate load impedance in the example, especially for the two lower bands. The coil taps are closer to the proper inductance for the three higher-frequency bands, but optimum efficiency and power output cannot be obtained unless the plate load impedance is more accurately matched for all five bands. The remedy is to rework the coil so that the taps are located to provide the inductances shown by the curves.

By consulting **fig. 1** and eyeballing the particular tapped coil under consideration (in this case the B&W 850A), it's fairly easy to judge where the taps should be relocated to provide the proper inductance for the particular plate load impedance involved. Clip leads may be used between the coil and switch points, after disconnecting the existing coil taps, to obtain the proper inductance experimentally. Then new taps can be installed in place of those on the as-built coil.

In my case, using a 4-1000A in grounded grid, I was able to increase transmitter power output substantially on all bands by relocating the taps on the B&W 850A coil in the amplifier pi network. If you'd rather not modify the coil, the clip-lead method can be used instead – a not too elegant way of doing the job, but the coil can easily be returned to its original configuration for trading purposes at the next swap meet.

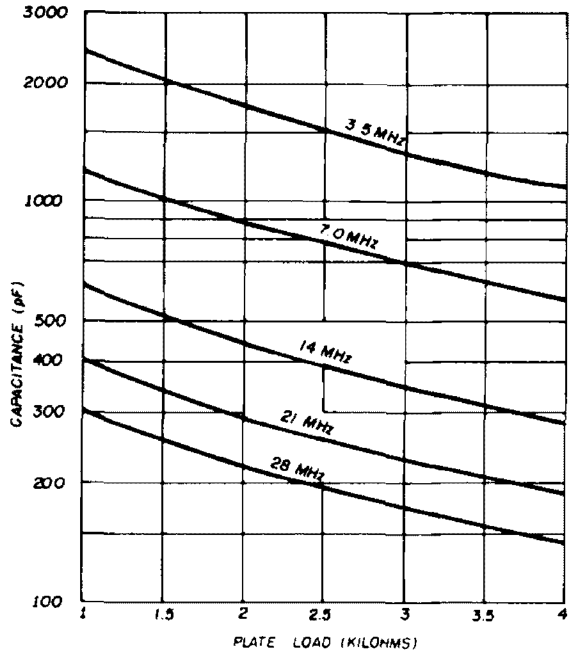

fig. 3. Pi network output capacitance as a function of plate-load impedance.

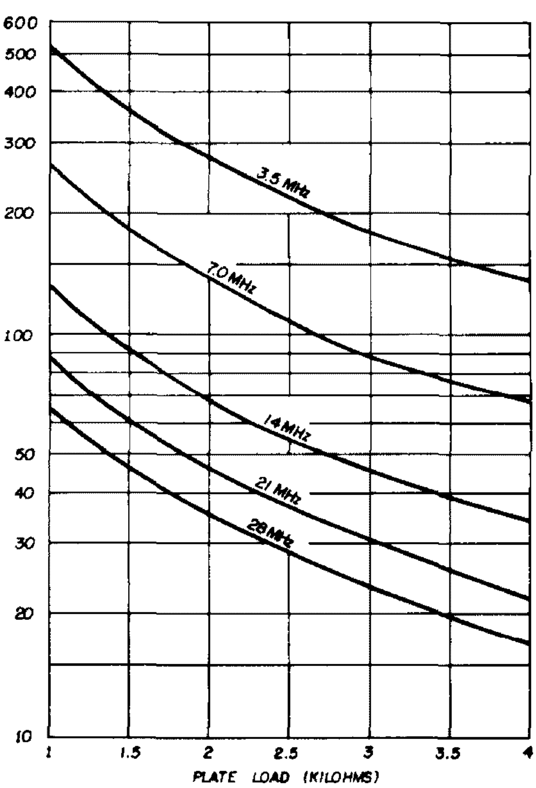

fig. 2. Pi network input capacitance as a function of plate-load impedance.

The curves in **figs. 2** and **3** are included for the remainder of the pi-network design problem. These curves provide values for the input and output capacitors (C1, C2) for plate-load impedances between 1000 and 4000 ohms. The same ground rules apply as for the inductor: operating Q of 12 and 50-ohm output impedance. You'll probably find that another compromise will be necessary in the choice of C2 for specific frequencies. For example, if you wish to use a variable capacitor for C2, it will probably be necessary to use fixed capacitors in parallel with C2 on certain frequencies to obtain optimum transmitter loading.

reference

1. Douglas A. Blakeslee, W1KLK, and Carl E. Smith, W1ETU/4, "Some Notes on the Design and Construction of Grounded-Grid Linear Amplifiers," *QST*, December 1970, page 22.

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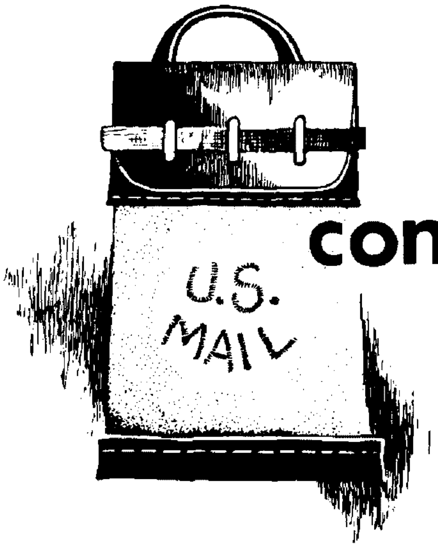


comments

log-periodic antennas

Dear HR:

I am afraid I have not stressed enough the importance of transposing the open center feedline that must be used with log-periodic antennas. It is an absolute *must*. Several amateurs who have written to me about log periodics built from my articles have indicated that the antennas have no gain but appear to have a bi-directional pattern off the sides of the antenna. In practically all cases this is caused by a failure to transpose the center feedline as shown in **fig. 1**. Note

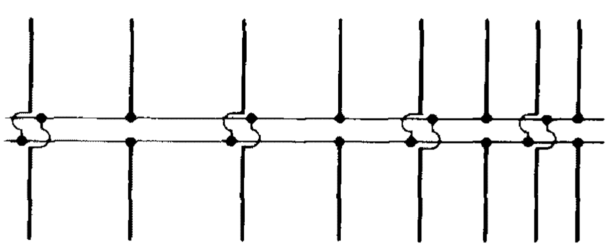


fig. 1. Center feedline connections to every other element of a log-periodic antenna must be transposed for proper operation, as shown here.

that every other element of the log periodic is connected to the opposite side of the center feeder. For more information on this essential part of log periodic construction, refer to my article in the September, 1972, issue of *ham radio* (page 33), or read my article on page 6 of this issue..

George Smith, W4AEO
Camden, South Carolina

Dear HR:

When I first built the five-element vertical 40-meter log periodic described by W4AEO in the September, 1973, issue of *ham radio* (page 46) and put it on the air, it didn't perform at all like it was supposed to. After checking all dimensions and the feedline transposition, WB4JEZ/5 and I sat down and scratched out heads — the antenna still didn't have any directivity. In a fit of desperation I drove a ground rod in at the balun, which is mounted at the front of the vertical array, and grounded the shield. What a difference! The antenna now works as advertised. The problem was a ground loop — without the additional ground rod the shield *at the balun* was not grounded.

Tom Morrison, WB5IZN

noise bridge

Dear HR:

After building the noise bridge that appears in the January, 1973, issue of *ham radio,* the following thoughts came to mind that might be useful for anyone attempting to build the noise bridge.

Although the author does not mention the wire size he used for winding the tri-filar winding on the T-37-10 toroid core, number-26 enameled worked very well. Once the primary side of the coil is connected to the circuit, the problem of finding the proper connections for the secondary with the four remaining wires can be solved as follows: wire this portion of the noise bridge last. With a receiver connected find the two wires that, when joined together and connected to the detector, will give the greatest noise output on the receiver (the remaining two wires must also be connected as per the schematic). Then connect a resistor of known value around 50 ohms to the Z_X

connector and adjust the R_X pot for a pronounced noise null in the receiver output. When this occurs, you know that the coil leads are correctly connected.

I might also mention that the 68-pF mica capacitor and the 140-pF variable capacitor were eliminated from my noise bridge since I am only interested in finding the resonant frequency of antennas, transmission lines, etc., and the resistance at resonance.

Since most amateur antennas have a resistance of less than 100 ohms at resonance a builder might want to substitute a 100-ohm composition pot for the 250-ohm pot used in the article. It was found on my noise bridge that the space between 10-ohm intervals was somewhat small when calibrating the resistance dial.

In comparing the performance of this noise bridge with an Omega TE-7-01 noise bridge that I have been using for three years, the results were very satisfactory. All tests were below 30 MHz.

John Lawson, K5IRK
Amarillo, Texas

standing-wave ratios

Dear HR:

The article in the July, 1973, issue by W2HB on "The Importance of Standing Wave Ratios" was very interesting and informative. Several small points seem, however, to invite some comment and clarification.

Figs. 1 and 2 are stated to depict conditions in a system with a vswr of 2:1. The curve shown in these figures is for a vswr of 3:1. Furthermore, its shape is shown as sinusoidal. When the magnitude of the voltage wave is considered without regard to polarity, which is the case here, the curve has the shape of a full-wave-rectified sinewave.

The first full paragraph at the top of the second column of page 32 regarding reflection at resistive terminations is somewhat misleading. It is certainly true that real power flows in only one direction — from generator to load. However, a reflected wave does exist at any resistive termination which does not match the impedance of the line. The implicit definition of "passive" ("That is, it cannot reflect . . . .") is not standard usage. Reactive elements may also be passive. This paragraph might well have been omitted from this otherwise generally very clear and useful article.

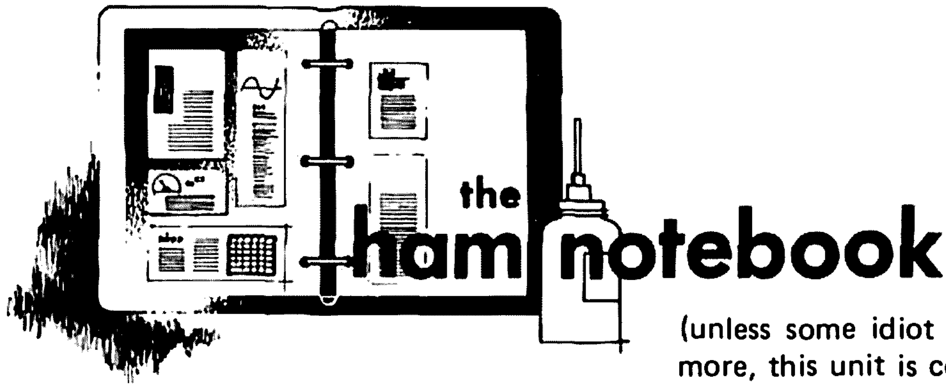
Kenneth H. Beck, W3VDX

Mr. Beck is quite correct in pointing out that the minimum voltage of the curves shown in my **figs. 1** *and* **2** *should be 0.5 instead of 0.3. This error is embarrasing for it resulted from having in mind the 3:1 ratio of the incident wave to the reflected wave that exists when the vswr is 2:1. Also, I have no quarrel with his comments about the shape of the curves. However, I would like to point out that insofar as the article is concerned, only the location of the peak and minimum voltage points is of importance, and the curve shapes shown should not be disturbing to the reader.*

Again I must agree with Mr. Beck that perhaps the choice of the word "passive" in discussing resistive loads was unfortunate because it is not standard usage. However, I was trying to stress the fact that a resistive load by itself does not, and cannot, reflect power in any kind of circuit including rf transmission lines. We consistently hear the statement made that power is reflected by the load. This statement is particularly incorrect when the load is pure resistance. Evidently the difficulties involved in explaining the mechanism of reflection has been avoided to the extent that the load has become credited with a false capability.

The mechanism by which reflection does take place on a transmission line has been described in excellent detail by W2DU in the August, 1973, issue of QST. *In this article the author has made it clear that reflection takes place as a result of the voltage and current conditions at the load, and the effect that these conditions have on the electric and magnetic fields at the load end of the line.*

Earl Whyman, W2HB



programmable calculator simplifies antenna design

Within the past year there has been a whole new series of programmable electronic calculators introduced to the market. These new calculators are a whole new breed of cat and are nothing like the home-variety hunt-and peck units that do a good job of wearing your fingertips down to the first knuckle. One of the programmable calculators currently available is the Compucorp 322G* which sells for slightly under eight bills. That's right, just this side of $800. For this price you get ten memory registers, 80 program steps and memory retention like a bear (unless some idiot turns it off). Furthermore, this unit is completely unbothered by large rf fields – at my station it sits placidly on the operating table, completely ignoring my pair of 813s.

Although the price tag is premium, you have to operate one of these programmable calculators to get the feel. It is all so simple. Just stick a few sample parameters into the memory registers and then load the program and switch the calculator back to *run*. Then you can play games with new items in the memory registers, hit the *start* button, watch it wink its readout light and hang up on the answer.

The Compucorp 324G programmable calculator used by the author.

antenna design

A programmable calculator such as the Compucorp 322G can be used in all kinds of antenna designs, but a simple two-tower directional antenna design will be presented here. However, there's no reason to stop at two verticals – the calculator will also handle four, provided you have the scratch to put up four towers. Nor is the calculator limited to antenna designs – it will handle nearly any sort of design problem you can load into it.

For example, you can load the calculator with a program to tell you which way to point the beam when you're working DX, and it will do the job quicker than your rotator can bring the beam around! I use a Compucorp 324G which stores two 80-step programs – the second to tell the guy on the other end which way to aim his beam.

For the specifics of the two-element directional array design, refer to fig. 1.

*Compucorp, manufacturer of the 324G programmable calculator shown in the photograph, is located at 12401 West Olympic Boulevard, Los Angeles, California 90064.

Compucorp®

Page 1 of 2

PROGRAM

DIRECTIONAL ANTENNAS, BASIC TWO TOWERS

Program No 8806242
Models 320 Series
Date 21 September 1973
Author Aylor

DESCRIPTION

This program calculates the radiated field intensity, from a two-tower directional array, at an azimuth angle, ϕ, measured from the reference axis (the line passing thru both towers) and at a vertical angle, θ, measured from the ground plane.

The program includes a stepping increment, allowing evaluation of field intensity for successive values of ϕ.

EXAMPLE:

Find the radiated field intensity from the following information:

- Azimuthal angle ϕ of 15°.
- Vertical angle θ, measured from ground plane of 0°.
- Field phase of second tower, A_2 of 35°.
- Ratio of field which second tower radiates above that radiated by reference element #1, R_2 = 1.15.
- Spacing between towers, S = 150.
- Multiplying Constant, K = 758.287 mv/m.
- Fill Constant, calculated from $(1.0 + R_2^2)/2R_2$, W = 1.00978.

Follow load procedure on reverse side. Steps 1-4, continue by:

ENTER	PRESS	READ	– Description
15	ST, 1		– Azimuth angle ϕ
758.287	ST, 2		– K
1.00978	ST, 3		– W
35	ST, 4		– A_2
150	ST, 5		– S
0	ST, 6		– Vertical angle θ
	S/S	74.997	– Radiated field int. in mv/m

Program No. 8806242
Page 2 of 2

OPERATING INSTRUCTIONS

1. Set switch to LOAD
2. Key in the program steps. (see note below)
3. Set switch to RUN, set DP to 3.
4. Set switch to DEG.
5. Store Registers 1-6 from table below.
6. Press S/S.
7. The radiated field intensity is displayed in millivolts per meter at one mile.
8. For additional increments, continue at Step 6.
9. For next problem, go to Step 5.

DATA STORAGE

Description	Symbol/Register
Azimuthal Angle, measured from ref. axis	ϕ/1
Multiplying Constant, field radiated from tower (2)	K/2
Fill Constant, $(1.0 + R_2^2)/2R_2$	W/3
Phase of field radiated by second tower	A_2/4
Spacing between towers	S/5
Vertical Angle, measured from ground plane	θ/6

NOTES:

1. The stepping increment is entered in the program starting at Step 36. Any desired value can be substituted to generate results for successive values of ϕ.
2. Multiplying Constant is calculated from [illegible]

REGISTER

0	1	2	3	4	5	6	7	8	9	10
	ϕ	K	W	A_2	S					

PROGRAM STEPS

Step	Key	Step	Key
1	PSET	41	
2	RCL	42	
3	1	43	
4	SIN	44	
5	F2	45	
6	×	46	
7	RCL	47	
8	5	48	
9	[illegible]	49	
10	[illegible]	50	
11	[illegible]	51	
12	RCL	52	
13	6	53	
14	SIN	54	
15	F2	55	
16	×	56	
17	[illegible]	57	
18	0	58	
19	+	59	
20	RCL	60	
21	4	61	
22	[illegible]	62	
23	[illegible]	63	
24	[illegible]	64	
25	[illegible]	65	
26	RCL	66	
27	3	67	
28	–	68	
29	SQRT	69	
30	×	70	
31	RCL	71	
32	2	72	
33	×	73	
34	[illegible]	74	
35	[illegible]	75	
36	[illegible]	76	
37	[illegible]	77	
38	[illegible]	78	
39	1	79	
40		80	

fig. 1. Compucorp's program for designing a basic two-tower directional antenna.

This is a reproduction of Compucorps's program 8806242 which is based on the design procedure described in Jasik's *Antenna Engineering Handbook*.[1] The program has a number of subtleties such as unequal ratios of rf fields between towers, vertical angle θ, which tells you what's happening at a specific radiation angle, and most important of all, a stepping increment which walks the thing around the azimuth by a specified increment (put in during the load phase) every time you hit the *start/stop* key. For the uninitiated, this is the old routine, "N = n + a," where N is the new value, n is the old value just presented and a is the increment.

The applications for the programmable calculator are limited only by your imagination. For those who are remiss to use their imagination, Compucorp will soak them $19.95 for the entire antenna design package.

reference

1. Henry Jasik, *Antenna Engineering Handbook*, McGraw-Hill, New York, 1961, equation 20-19.

Raymond Aylor, W3DVO

noise bridge

The prospect of tuning a new quad and adjusting a shunt-fed tower for 160 meters led to the construction of the rf bridge described by WB2EGZ (*hr*, December, 1970, page 18). The completed instrument performed perfectly, with very deep calibration nulls. However, once the bridge was connected to an existing 20-meter beam to determine that antenna's resonant frequency, a serious flaw appeared — signals on the ssb segment of the 20-meter band were anywhere from 5 to 30-dB stronger than the noise from the bridge.

W5QJR mentioned the need for high-level amplification of the noise generated by the bridge in the first article on this type of instrument several years ago ("The Antenna Noise Bridge," *QST*, December, 1967, page 39), and I learned quickly that this was the case. A stage-by-stage check revealed that everything was functioning properly, but that gain

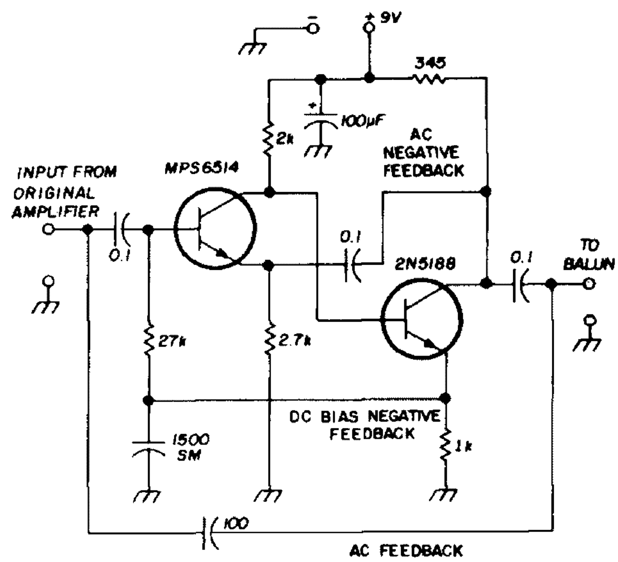


fig. 2. Broadband noise amplifier for use with antenna noise bridges for high-frequency measurements.

through the broadband amplifier was on the order of a mere 15 dB. Not quite enough for normal signal conditions on the high-frequency bands.

The circuit of fig. 2 was breadboarded and found to provide 35 to 50-dB of additional gain (not entirely constant over 1.8 to 30 MHz). Three strong feedback loops are introduced between the driver and final amplifier. Both devices are high f_T, high beta types, and substitutions of other popular devices (such as the 2N3053, 2N2102, 2N697, 2N706, etc.) will cut overall gain by 10 to 20 dB, so the devices specified should be used. The amplifier was mounted on its own PC board, which in turn was mounted in the Minibox on the opposite side from the first PC board (where the battery is shown in the WB2EGZ instrument).

The battery fits snugly between the ends of the PC board and the bottom of the Minibox. Lead wires were simply run from the output of the WB2EGZ amplifier to the input of the additional amplifier, and the output from the second amplifier was connected to the broadband balun. One slight improvement over the WB2EGZ version involves mounting the potentiometer terminals so that they point directly to the input/output receptacles, rather than as shown in the photos. Zero lead length is thus achieved by mounting each component of the bridge directly to the receptacles and the pot terminals. No reactance arises as a result.

The additional broadband amplifier provides a noise signal that blankets any signal on the high-frequency bands. Further, accuracy of null detection is greatly increased since the noise drops about 40 dB to reach incoming signal levels, and a complete null is down about another 10 to 15 dB. I've found that the bridge now can be tracked down to about a 5-kHz bandwidth.

Ade Weiss, K8EEG

ST-5 keys polar relay

Here is a very simple way to key a polar relay with the ST-5 RTTY demodulator (see fig. 3). I use this method to key my transmitter and it works very well. It may also be used to key an AFSK oscillator which feeds a tape recorder. This way you can make a recording on the hf bands and replay it on vhf.

Fred Gilmore, WØLPD

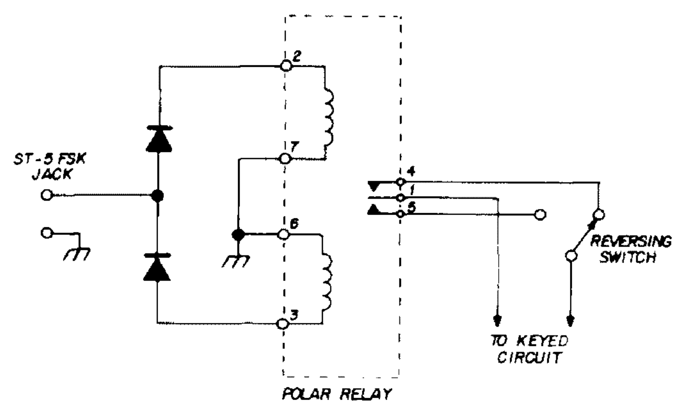


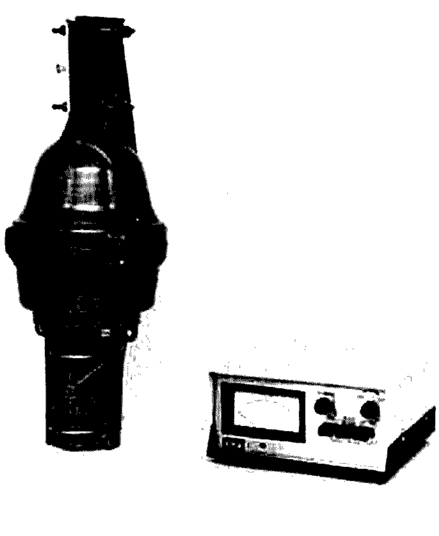
fig. 3. Simple circuit for keying a polar relay with an ST-5 RTTY demodulator.

short circuit

In the article describing the Universal Frequency Standard in the February, 1974, issue, the HEP50 in fig. 5 (page 44) is missing a 1000-ohm collector resistor; the printed-circuit layout on page 46 is okay. Also, the wiper of PA-12 (switch S2) should be grounded. The output from pin 11 of U5, a 7493 binary-counter IC, is 12.5 kHz, *not* 2.5 kHz.

new products

new ham-m rotator



The new Ham II rotator introduced recently by CDE succeeds the popular Ham-M rotator used by amateurs for years. The new Ham II features a new brake release control, separate directional controls and stainless-steel gears and hardware.

The all new, modern design control box is styled to compliment surrounding communications equipment and provides built-in, operator-controlled brake release for improved longevity of the entire antenna/rotator package. The calibration control is now located on the front panel for ease in maintaining directional accuracy. Also included on the front panel is a separate off/on switch which offers continuous direction indication on the illuminated meter, making access to the rear of the control box unnecessary. Individual snap-action switches are used for rotation-direction control.

The new Ham II rotor continues the tradition of the heavy-duty, cast-aluminum bell housing, long the trademark of CDE amateur antenna rotators. The inline construction evenly supports the load on two six-inch bearing races containing 98 ball bearings. An electrically-controlled wedge brake is housed in the base, positively locking the rotor in any of 96 segments (3° 45" apart). The high torque motor drives the unit through a machined stainless-steel gear and pinion assembly, rotating a full 360 degrees in less than 60 seconds. Designed for antennas of up to 7.0 square feet of wind load area, the rotor accepts masts from 1-3/8 to 2-1/6 inches (3.5 to 5.2 cm). A tower-mounting kit and south-center meter-scale kit are available.

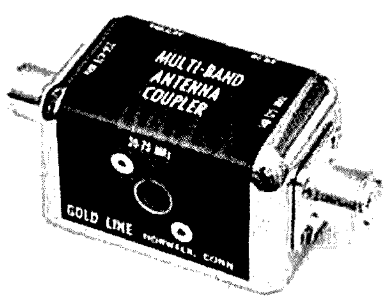
Also new from CDE is the CD44 antenna rotator which succeeds the popular, intermediate range TR44 often used by amateurs for smaller antenna systems. For more information on the new Ham II or CD44, write to Cornell-Dubilier Electronics, Division of Federal Pacific Electric Company, 2070 South Maple Street, Des Plaines, Illinois 60018, or use *check-off* on page 110.

two-meter colinear

The new Hustler two-meter colinear antenna, Model G6-144-A, is expressly designed for repeater or fixed station operation on two meters. FCC accepted for repeater application, the antenna is

conservatively rated at 6-dB gain based on EIA Standard RS-329 (gain over a 1/2-wave dipole). Special features built into this 117-inch antenna include high-power capability, shunt feed with dc grounding, easily accessible SO-239 coax connector, four radials, heavy duty construction and double U-bolt mounting. Price is $49.95 amateur net. For additional information, write to Sales Department, New-Tronics Corporation, 15800 Commerce Park Drive, Brookpark, Ohio 44142, or use *check-off* on page 110.

multi-band antenna coupler



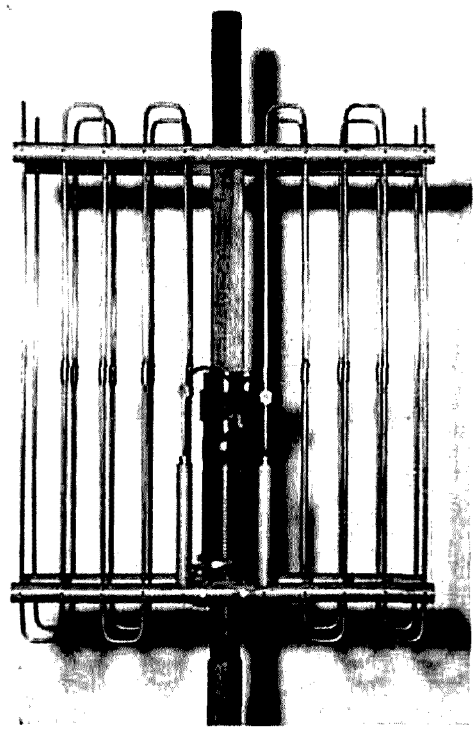
The new Gold Line GLC 1079 Multi-Band Antenna Coupler allows you to use your standard car radio antenna to monitor 20–70 MHz, 148–175 MHz and 250–470 MHz as well as your a-m/fm car radio. It can couple up to five bands without a new or special antenna. Two cables are included for easy hook-up. Price is $12.95 from Gold Line Connector, Inc., Muller Avenue, Norwalk, Connecticut 06852. For more information, use *check-off* on page 110.

low-loss uhf coax

The Antenna Specialists Company has introduced what it claims to be the first major development in communications coaxial cable in twenty years, PRO-FLEX$_{tm}$ 450, which offers significantly reduced power loss in the ultra-high-frequency range and higher ambient temperature ratings. The new cable, which is intended primarily for vehicular

installations but may also be used effectively in base station installations of moderate power, is rated at 150° F ambient rather than 104° F commonly associated with conventional RG-58/U cable. The cable construction consists of an inner-conductor, an inner-dielectric of foam polypropylene, shield braid and an outer jacket of purest non-migrating vinyl. Antenna Specialists is now converting production of its heavy duty uhf and vhf professional mobile antennas to include the new PRO-FLEXtm 450 cable. It also is available in bulk quantities of 100, 500 and 1,000 feet for the convenience of systems designers and for replacement applications. Complete specifications may be obtained by writing to: Mr. Larry Kline, Professional Products Manager, The Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106, or by using *check-off* on page 110.

little giant antenna

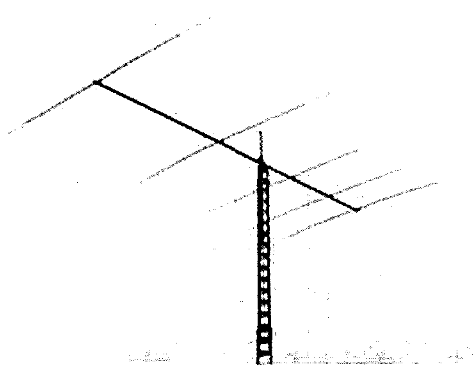


Stan Byquist's "Little Giant" antenna is not exactly new—he received his patent on it back in the 1950s and there was some publicity for it then—but he has recently resurrected the design and its unique character deserves mention in these pages.

More Details? CHECK—OFF Page 110

In brief, the Little Giant is a highly compressed single-band antenna that can be used in situations where conventional antennas are out of the question. The largest model, for 80 meters, is only 27-inches (68.6-cm) high and 32-inches (81.3-cm) wide; even smaller models are available for use on 40 through 10 meters. Bandwidth is necessarily small, as would be expected of such a small and, therefore, high-Q antenna. User reports have been quite favorable considering that any drastically shortened antenna is bound to be a compromise in performance. Amateurs with a space problem should contact Stan at the Little Giant Antenna Labs, Vaughnsville, Ohio 45893, or use *check-off* on page 110.

monobander antenna



KLM Electronics has introduced a new, 20-meter "big stick" monobander antenna which provides constant gain *and* flat vswr over the complete frequency range from 13.9 to 14.4 MHz. By using very efficient driven elements, the five-element array provides optimum gain of 9.7 ±0.2 dB gain over a dipole with better than 35-dB front-to-side and 30-dB front-to-back ratio. A boom length of 41 feet and turning radius of 28 feet are combined with rugged construction to yield a beam that weighs in at only 65 pounds. Feed impedance is 200 ohms, or 50 ohms with KLM's optional 4-kW PEP balun, priced at $14.95. The 20-meter monobander is priced at $199.95. For more information, use *check-off* on page 110, or write to KLM Electronics, 1600 Decker, San Martin, California 95056. If you're in a hurry, call them at (408) 683-4240.

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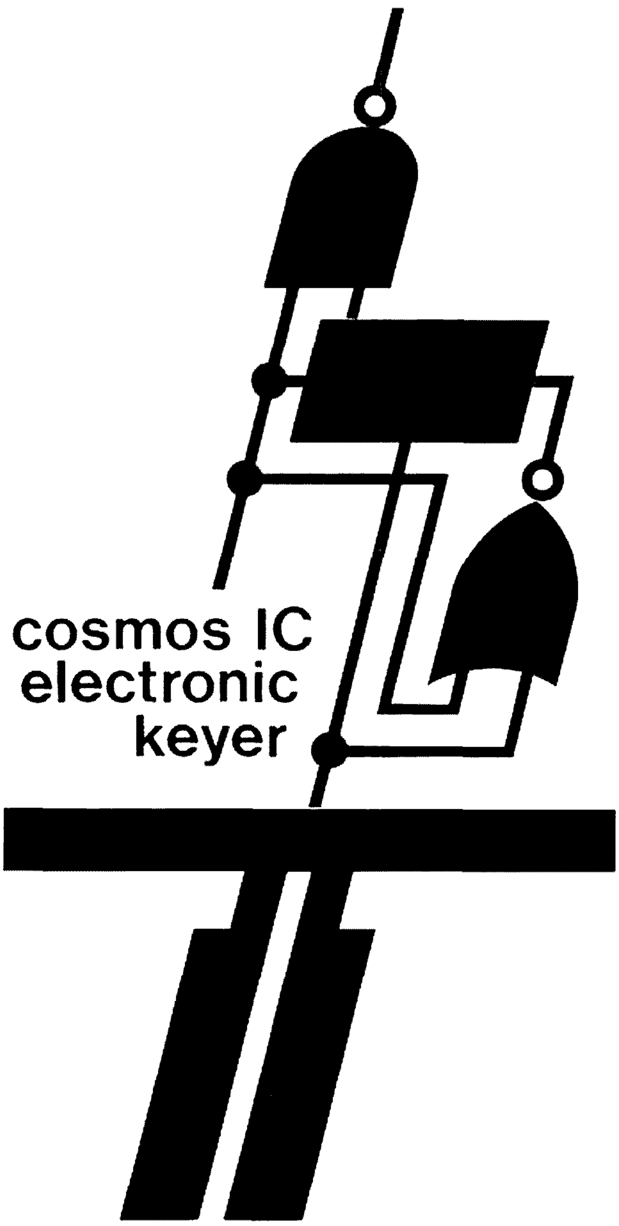
ham radio magazine

JUNE 1974



cosmos IC electronic keyer

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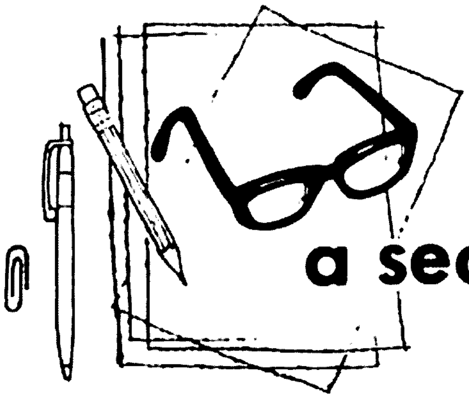


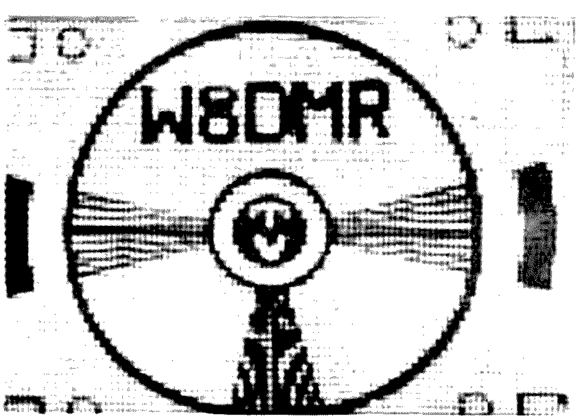
a second look

by jim fisk

Solid-state imaging devices which may eventually find their way into high-quality television cameras are now becoming available for lower-resolution applications such as slow-scan optical character recognition and basic-recognition security systems. Some of these devices, which consist of large arrays of charge-coupled photo-sensitive semiconductors that are scanned with digital circuitry, are so sensitive they can detect objects illuminated only by candle-light. The big problem has been to obtain the 525-line video resolution required for television cameras. However, Bell Labs has built a solid-state vidicon capable of full *Picturephone* resolution (250 by 225 lines) and Fairchild has a 100-by-100 element imager that is suitable for many industrial applications.

Actually there are two different types of solid-state imaging devices: the charge-coupled device or CCD and the charge-injected device or CID. Although the image signal is developed and stored in both devices in much the same way, the signal is retrieved differently. In the CCD array the image is scanned sequentially while in the CID array the image is scanned in an x-y fashion. This permits the CID array to be scanned at any speed or individual points to be scanned randomly. Another advantage of the CID design is that single-image defects cause the loss of only one image point. In a CCD array a single-image defect blanks out the rest of that line.

ATV test pattern from W8DMR is generated by solid-state charge-coupled image sensor.



Bill Parker, W8DMR, is believed to be the first amateur to use a solid-state image sensor to transmit an amateur television test pattern. Bill, who has been on amateur television since 1949 and is still active, uses the 10,000-element Fairchild CCD-201 solid-state imager mentioned earlier. This sensor is mounted in a 24-lead dual-in-line package with an optical glass window. The 1.2- by 0.8-mil (0.03- by 0.02-mm) image-sensing elements are located on 1.2-mil (0.03-mm) vertical centers and 1.6-mil (0.04) horizontal centers. This provides an image aspect ratio of 4 by 3. In addition to the image-sensing chips the CCD-201 includes 100 columns of two-phase shift registers interdigitated among the photo-sensitive elements, a 102-unit two-phase analog output shift register, an output preamplifier and a compensation amplifier. A photograph of the transmitted test pattern is shown to the left. When 512-by-320 element arrays become commercially available the resolution wedges on the test pattern will be much more distinct. Who will be the first to apply this new technology to real-time slow-scan TV?

Jim Fisk, W1DTY
editor-in-chief

cosmos IC electronic keyer

Applying cosmos technology to a versatile, compact, low-drain keyer design

James W. Pollock, WB2DFA, 6 Terrace Avenue, New Egypt, New Jersey

Since the advent of ICs on the amateur radio scene, a number of articles on keyers have appeared in the amateur magazines. Now that c-mos or cosmos ICs have become available to the amateur from various sources at modest prices,* it's time that these "state of the art" building blocks found an application in a homebrew project.

The cosmos keyer described here draws only 0.4 mA on standby, with an average key-down current drain under 2 mA at a supply voltage of 10 volts. The keyer can work properly with supply voltages from 4 to 15 volts. When operated at 5 volts, the keyer actually consumes less than 100 microamps, less power than a set of headphones! Its low power requirement makes it ideal for the QRP or field day enthusiast. If TTL logic was substituted for the cosmos ICs in this keyer, the current drain would be in excess of 200 mA.

Some of the more important features of an electronic keyer using cosmos ICs is low power consumption, simple construction and modest construction cost (about $10.00 for parts). The keyer described in this article features self-completing dots, dashes and spaces, a built-in transmitter keying circuit and sidetone generator and small size — the circuit board can be mounted inside your favorite rig.

*Poly Paks, Inc., Box 942, South Lynnfield, Massachusetts 01940.

The basic clock gating circuit used in the keyer is illustrated in fig. 1. The clock gate allows only full clock pulses to pass through it, regardless of the timing inaccuracies of the *enable* signal. No partial pulses or "slivers" can be tolerated if a keyer is to send perfect code.

In fig. 1, U1 is a type-D flip-flop. The logic level present at the *data* or D input will be transfered to output Q at the next positive going edge at the *clock* input (pin C). Notice that the clock feeds pin 1 of U2, pin 1 of U3 and pin C of U1. U2 is the output gate and U3 is the reset gate.

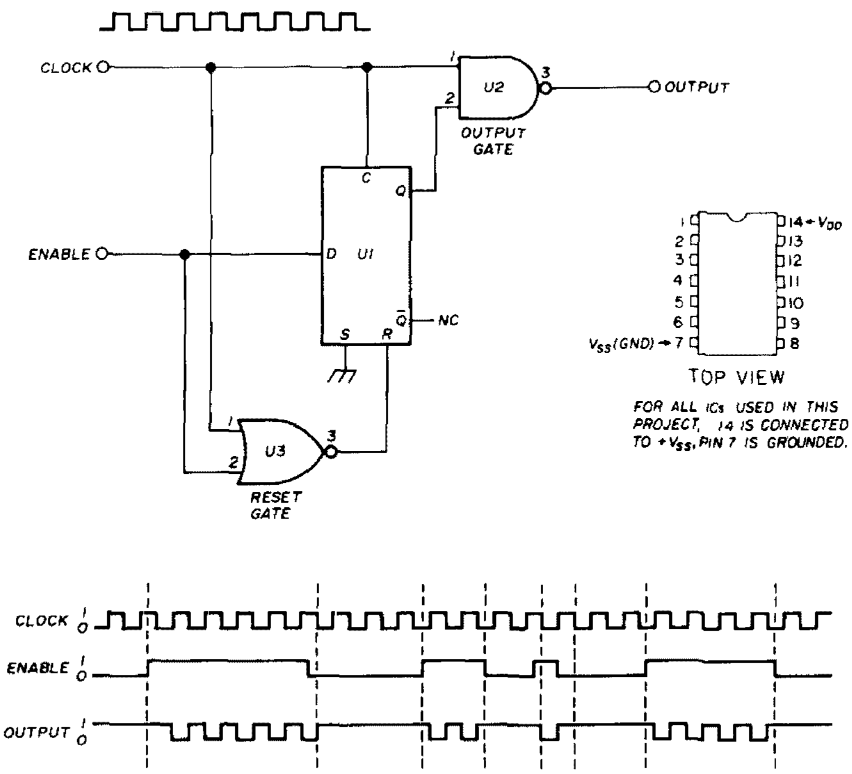


fig. 1. Basic clock gating circuit for the cosmos keyer.

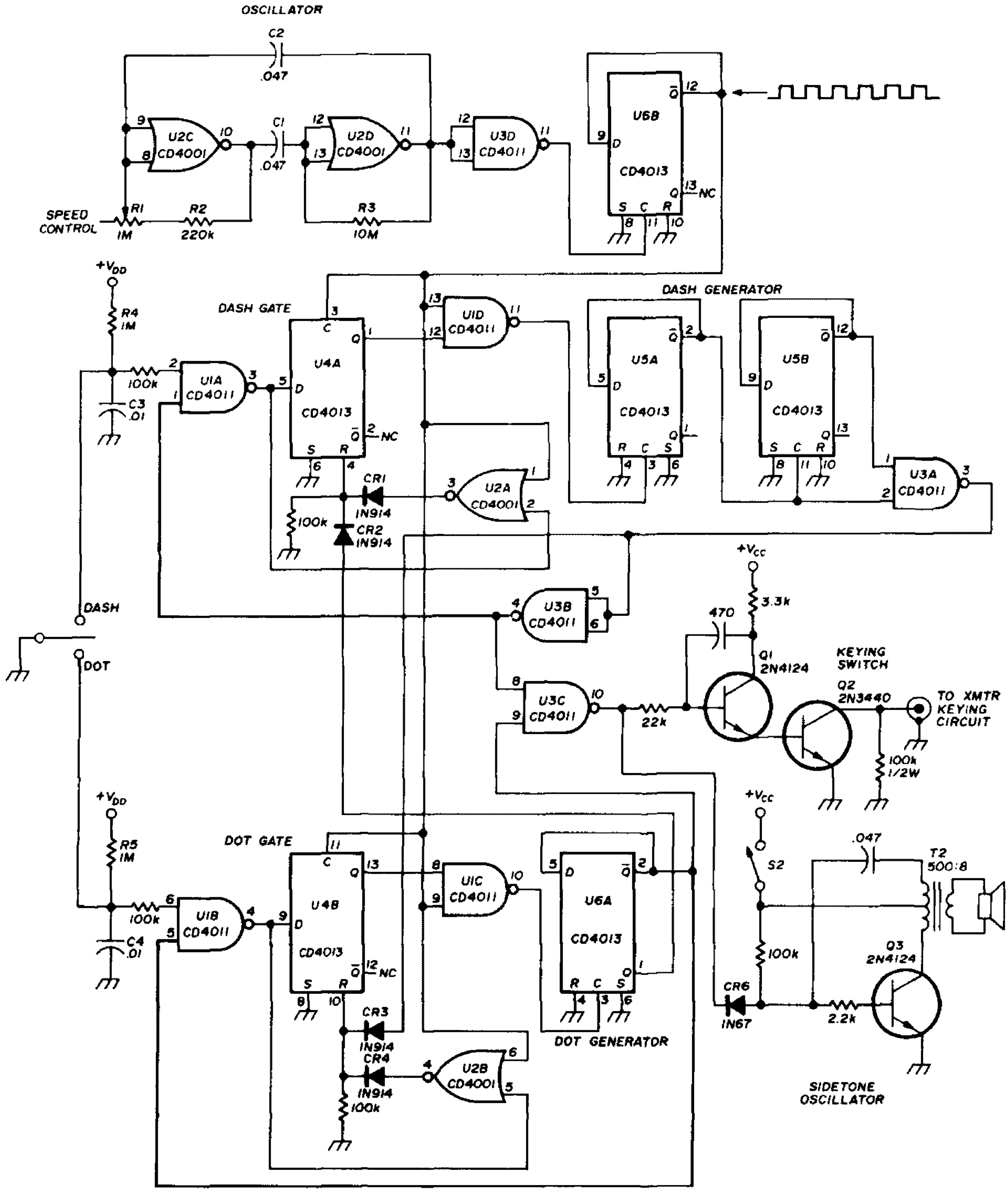
When the *enable* signal is low (zero), pin Q of U1 is assumed to be zero if pin C is being clocked. As long as pin Q of U1 is zero, U2 will not pass the clock pulses. Also, as pin D of U1 is zero, the clock pulses at pin 1 of U3 will pulse the reset pin (pin R) of U1. A 1 on the reset pin of a cosmos flip-flop will force Q to zero and will override all other input signals. The *set* pin of U1 is grounded for this reason. A 1 at the *set* pin will force Q to a 1 and will override all other input signals.

When the *enable* signal goes to a 1, pin Q will go to a 1 at the next positive-going clock pulse leading edge. Now that pins 1 and 2 of U2 are 1, pin 3 goes to zero and will go to a 1 when the negative-going clock transition occurs. Thus the output of U2 is the inverted version of the clock pulses.

When the *enable* goes to zero, flip-flop U1 will be reset by the NOR gate U3 at the time when the clock is zero. If the NOR gate were not used to reset U1, then the next positive-going edge of the clock pulse would appear at the pin 1 of U1 simultaneously with the negative transition at pin 2 of U2 due to the data transfer action of flip-flop U1. Thus you would have logic levels changing simultaneously in opposite directions at pins 1 and 2 of U2. This would result in a "sliver pulse" at pin 3 of U2 when the input signals cross the threshold level of U2. To prevent this from happening, the reset gate U3 is used to reset the flip-flop when both the *enable* and *clock* inputs are at zero.

final circuit

The circuit I finally settled on for my cosmos keyer is shown in fig. 2. The keyer generates dashes and dots at a fixed time ratio of 3 to 1. A space has the same duration as a dot. The time base of the keyer is generated by two NOR gates, U2C and U2D, connected in a class-A multivibrator configuration. Resistor R3 causes U2D to self bias into a class-A condition, with the output reaching a dc level equal to the threshold of the gate itself (about 45 percent of the supply voltage). Resistors R1 and R2 have the same effect on U2C. The time constant (R3-C1) is much greater than (R1 + R2)C2 so the frequency of the oscillator

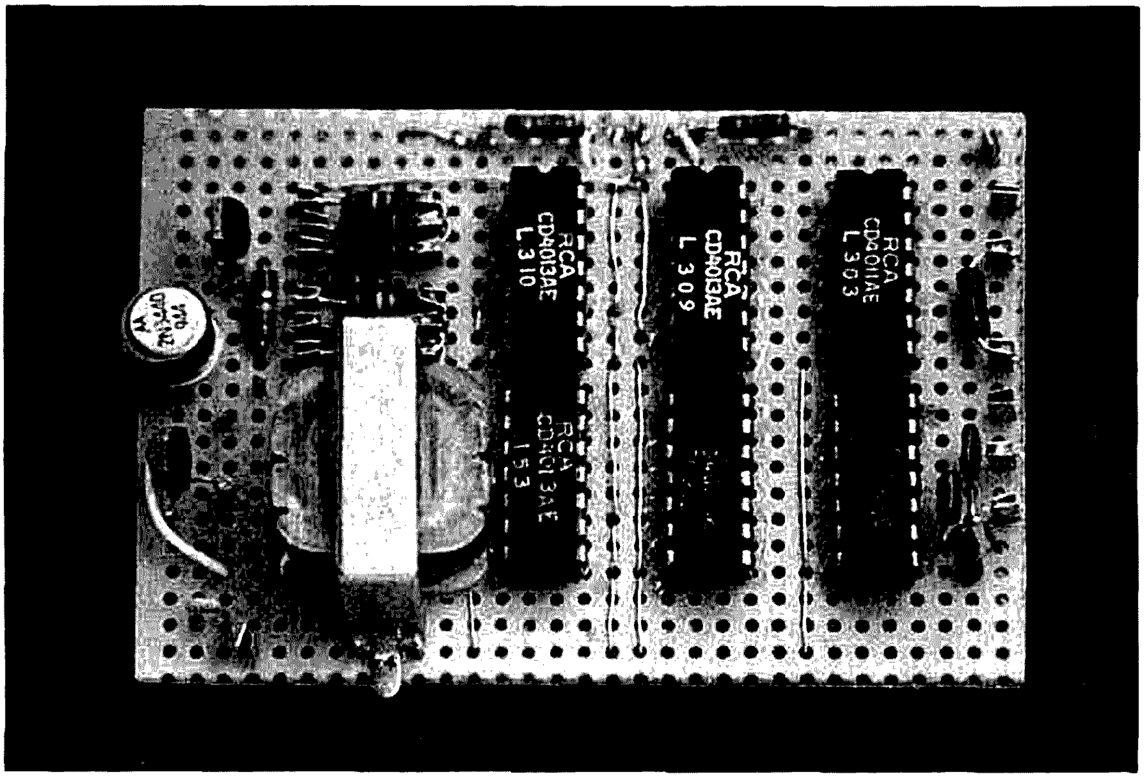


R1	1 megohm linear taper potentiometer	U2	cosmos NOR gate (CD4001 or equivalent
T2	500 ohm CT to 8 ohm transformer (Radio Shack 273-1381)		
U1, U3	cosmos NAND gate (CD4011 or equivalent)	U4, U5 U6	cosmos dual D flip-flop (CD4013 or equivalent)

fig. 2. Complete schematic diagram for the cosmos keyer. The bold lines are the feedback paths for the dot and dash generators, which allow them to end their timing in sync with the clock after paddle release. The sidetone generator, lower right, is optional.

is inversely linear with the setting of R1. Inverter U3D buffers the oscillator and squares up its output. Flip-flop U6B divides the oscillator frequency by two, but more importantly, provides a clock source with a perfect 50 percent duty cycle.

Note that the dot and dash generators each have their own clock gates, and are connected in such a manner that which-



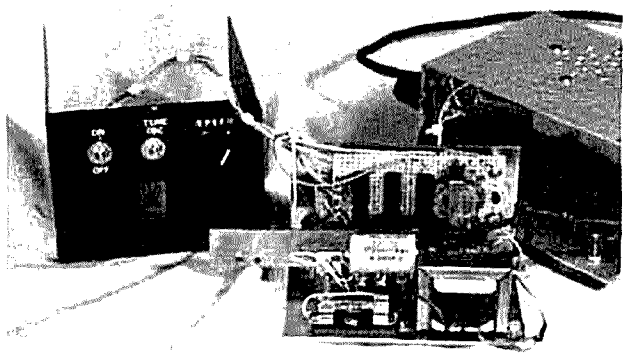
Logic circuitry for the cosmos keyer is wired on small section of perforated circuit board. Keying transistor and sidetone circuit are to the left.

ever of the two gates is enabled first overrides the other until its timing cycle is completed. If the dot gate is enabled first, the dash gate will be held in reset via diode CR2 until the timing cycle for the dot is completed, even if the dash paddle was depressed. Also, a complete space period would elapse before a dash could be sent, and vice versa. Diode CR3 allows the dash gate to reset and override the dot gate while dashes are being sent. The RC networks R4-C3 and R5-C4 provide pull-up bias for gates U1A and U1B and also eliminate the effect of key contact bounce.

The dash generator consists of a four-state binary counter (U5A and U5B) and a gate (U3A) to decode the four count states into dashes that are three clock periods long, separated by spaces one clock period long. U3B inverts the output of U3A to provide the proper logic levels to U3C by cancelling the inverting effect of U3A.

The bold lines are the feedback paths for the dot and dash generators. These feedback lines allow the dot and dash generators to end their timing in synchronization with the clock after their respective paddles have been released. This is what adds the self-completing feature to the keyer.

The output gate U3C drives the Darlington pair Q1 and Q2. Transistor Q2 is a high-voltage device, the Motorola 2N3440, which has a 250-volt B_{Vcb} rating. If you plan to use the keyer solely



Construction of the cosmos keyer. Circuit board, keying mechanism and power supply are on center chassis. Speed control and switches are mounted *on front panel, left.*

for your solid-state QRP rig, then just about any 5-watt npn transistor will do the job.

The largest portion of the standby current drain can be attributed to the oscillator. When the threshold region of a cosmos device is entered at a slow rate (such as by RC decay), the device draws relatively large amounts of current. The increase in current during key-down conditions is caused by the conduction of Q1 in driving Q2.

important

When working with cosmos ICs it's important to remember to *never* leave unused input pins floating. *Always* make sure that any unused input pin is tied to either ground or to V_{dd}, whichever is logically appropriate. For example, the *set* and *reset* pins of U5 and U6 have been tied to ground. If this precaution is not observed, these high impedance inputs are wide open for electrostatic charge pick up. Also, since the input capacitance of a cosmos device is typically 4 pF and the gate impedance is on the order of 10^{12} ohms, the result is a parasitic RC network with a time constant of 4 seconds. Any electrostatic charge can be stored for several seconds, injecting a false logic level into the cosmos device. That could raise havoc with your logic.

If you follow the schematic faithfully, all of the cosmos gates and flip-flops will be used up in fabrication of the keyer, with no surplus devices to cause problems. *Check* and *doublecheck* your wiring!

It's also advisable to provide overvoltage protection and regulation if operation from an unstable supply is anticipated. *Do not* exceed 15 volts, or

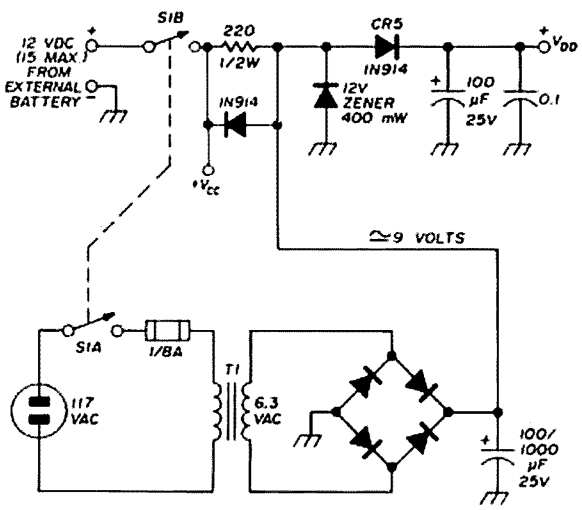
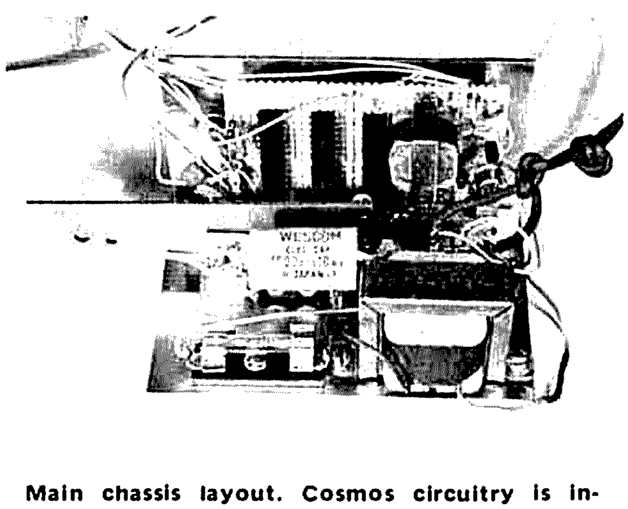


fig. 3. Suggested power supply for cosmos keyer, with inputs for either 117 Vac or low voltage dc. Transformer T1 is a small 6.3 Vac filament transformer.

the cosmos ICs will zener and draw excessive current which may destroy them. I recommend a maximum supply voltage of 12 volts. This should give you plenty of margin for error. Also, be sure to provide a means of protection from accidental polarity reversal of the power supply. Diode CR5 in my supply, fig. 3, provides this protection.

When soldering cosmos ICs into a circuit, use an iron with a grounded tip. Finally, make sure that your paddle key is clean. Any leakage path to circuit ground that is less than one megohm will falsely trigger the input gates.

The sidetone generator is optional, but can be included if your present rig doesn't have one. The sidetone adds about 3 mA to the key-down current drain.

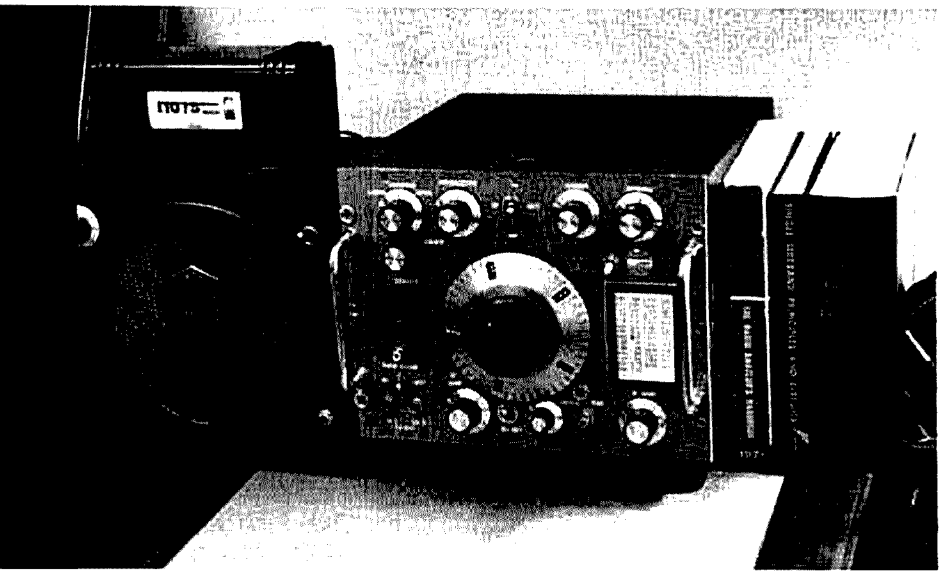


Main chassis layout. Cosmos circuitry is installed on perforated board, rear; power supply is in foreground.

reference

1. *COSMOS Digital Integrated Cricuits*, RCA Data Book Series, SSD-203A, RCA Semiconductor, 1973.

ham radio



some design ideas for specialized communications receivers

If you have ever considered building your own communications receiver, here is a collection of design ideas you may want to use

Ray Moore, 116 Walnut Street, Walpole, Massachusetts

When I read the editorial in the June 1972, issue of *ham radio* I felt like thı boy who, after Shoeless Joe Jacksoı confessed his involvement in the 191! Black Sox scandal, said, "Say it ain't so Joe." In that editorial Jim Fisk declared " . . . we at *ham radio* note that the da) of the amateur-built receiver may havı passed in favor of the vastly superior anc less expensive commercial version."

Say it ain't so, Jim. Yet, in one sense, a least, the statement is true. The individua amateur can no more hope to beat thı professional receiver designer at his owr game than he can hope to build a bettei and cheaper family sedan in his back yard.

The catch is that the professiona receiver designer's "game" is *not* to de sign the ultimate performance set but tc design a mass-market, multiband, multi mode, decorator-styled receiver that car sell at a popular price. Many of thı homebuilt receivers appearing in the harr

magazines are imitations of these commercial designs and, while their builders have the statisfaction of gaining valuable experience and operating homebrew equipment, the result is likely to be an inferior receiver at a higher cost.

Just as there are cars which will outperform the best that Detroit offers, so there are receivers which will outperform any amateur receiver on the market. Free the designer from the price restriction and you get cars like the Jaguar and Rolls-Royce and receivers like the National HRO-600 and the Collins 651-S1. If the product can bypass the mass-market criteria, if it can be designed for a specific, limited purpose, then, as W8YFB has said,[1] any one of 100,000 teenagers can build a performance car that will outperform Detroit's creations on the drag strip. And many an amateur can build a better receiver than he can buy.

Goodman made the point back in 1951, " . . . the fellows who build their entire receivers from scratch . . . don't do this because they can't afford a commercial job – often they . . . could afford several such receivers – they do it because they know exactly what they want and this is the only way to get it."[2]

Let's look at some of the restrictions imposed on the professional designer of amateur communications receivers. First, the company's management and marketing people place a selling price of, for instance, $400 on the receiver. Unfortunately, advertising, sales costs, shipping, warranty costs, taxes, engineering, overhead and manufacturer and distributor profits eat up $250 of the selling price. This leaves less than $100 for the cabinet and all the parts which go into it. Finally, only $50 is left for all direct labor on the receiver, assembly, alignment, testing and troubleshooting.

The receiver must cover all modes and the six bands from 160 thru 10 meters with single knob bandswitching. It must fit a low-profile, decorator-styled cabinet that matches the other equipment in the company's line. All components must be readily available in quantity and performance must be competitive with other receivers in the same price class.

The homebuilder, on the other hand, can design his receiver for specific objectives. He can design a "performance" or "competition" receiver for a particular band and activity. How about an all-out receiver for 160-meter CW DX or 20-meter ssb DX or 20-meter sstv or 40-meter RTTY or for 20-meter operation in downtown Los Angeles? Net operators could have a switch-tuned, maximum convenience job. The traveling man can design an outstanding, ultraportable receiver.

The limited versatility of such receivers is justified because the amateur may spend 80% or more of his spare time enjoying his particular operating specialty. His chrome-plated commercial job then becomes a second receiver for the occasional excursions to other bands and modes. Sometimes the homebuilt "performance" receiver makes an excellent tunable i-f for other bands.

When designing your receiver at home you need consider only your own interests and radio environment. You can use one-of-a-kind, high quality surplus parts that commercial people can't afford. You can add extra parts and extra stages – if a commercial designer eliminates a $5 stage he saves his company $50,000 over a 10,000 unit production run, and employers look with favor on people who put $50,000 back into the company's pockets. Furthermore, you can use circuits that require considerable diddling and alignment, and you can use the latest components and techniques without worrying about obsoleting a large inventory of parts.

One of the delights of home building is that you can over-design. Instead of using the flimsiest possible chassis material you can start out with a 1/8-inch aluminum plate. You automatically insert parasitic suppressors in all amplifier and oscillator

input and output elements. Where the professional runs endless tests to eliminate a capacitor here, a resistor there or a shield somewhere else, the home-builder can double decouple every stage and put in extensive shielding. Where the professional must run two i-f stages on the verge of oscillation, you can run three that loaf quietly along. Space, band-switching and ganging considerations need not limit the number and Q of the tuned circuits used in the front-end.

The receiver shown in this article, the DBR-1, illustrates many of the advantages the homebuilder enjoys. Its primary purpose is to provide superior reception of foreign a-m stations on the standard broadcast band in the presence of interference from the many domestic broadcasters. Secondarily, it is an excellent tunable i-f for shortwave broadcast and amateur ssb reception. While few amateurs are interested in broadcast-band DX, this receiver includes many techniques and ideas that are applicable to homebuilt receivers in general.

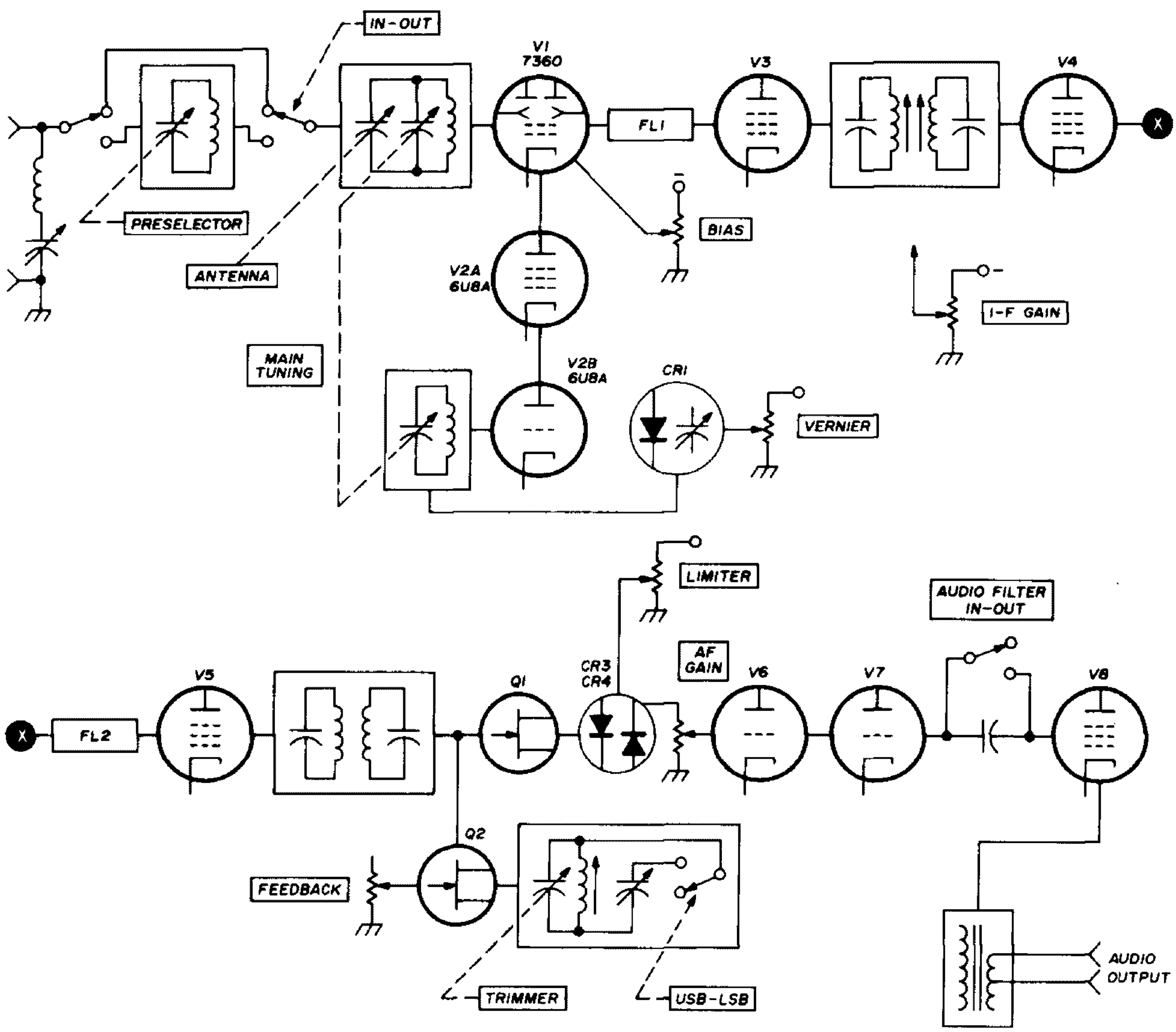


fig. 1. Block diagram of the DBR-1, a specialized DX receiver.

homebrew receiver

Vacuum tubes predominate in the DBR-1 because it was started back in 1967 when tubes had performance advantages in many circuits. Referring to fig. 1, the first circuit is a series-tuned i-f trap to eliminate feedback problems en-

countered at high i-f gain levels with no rf stage.[3] Next is an additional preselector circuit which can be switched in to eliminate images when using a long-wire antenna. A tuned loop antenna is normally used with the receiver and then images are not a problem.

The first tube, V1, is a 7360 mixer. The vfo and buffer amplifier is a 6U8A. A varactor diode, CR1, provides vernier tuning of the oscillator to facilitate precise positioning of the carrier for exalted-carrier a-m reception. Bias for the 7360 mixer is adjustable from the front panel for experimental purposes.

A 455-kHz i-f is used to take advantage of the sharp and inexpensive Kokusai mechanical filters. In accordance with good strong-signal practice, the first filter is placed as close to the antenna as possible, at the output of the mixer.[4] A second mechanical filter, between the second and third i-f stages, combines with the first to give a stop band over 120-dB down and a shape factor approaching 1:1. Three i-f stages (V3, V4 and V5) are necessary to provide sufficient gain to make up for filter losses and the lack of an rf amplifier.

Transistor Q1 is an infinite impedance detector; Q2 is a Q-multiplier to provide the peak in the i-f passband which exalts the a-m carrier. A switch changes the peak from one side of the passband to the other for receiving either USB or LSB. The feedback control adjusts the height of the peak and thus the amount of carrier exaltation. Diodes CR3 and CR4 are a full-wave audio limiter.

The gain of the receiver is kept as low as possible through the detector, and three stages of audio (V6, V7 and V8) are required to bring the overall gain up to an acceptable level. A small coupling capacitor to the output stage can be switched in to provide low frequency roll-off. A high-impedance output for a tape recorder and a low-impedance output for stereo phones or a speaker are also provided.

An OB2 voltage regulator, V9, regulates the oscillator plate voltage and Q-multiplier drain voltage. A zener diode, CR2, also works from the +105-volt regulated line to provide additional regulation for the varactor tuning diode. The power supply is in a separate cabinet. It uses five silicon rectifiers and supplies 6.3 Vac for the heaters, a 250-volt B+ supply controlled by an auto-transformer and a 0-150-volt bias supply.

mechanical details

Here is where most receivers are inadequate. They are flimsy, using the lightest possible material for the chassis and cabinet. The DBR-1 has a 1/8-inch aluminum panel bolted to a 1/8-inch chassis plate which is then bolted into a rugged, welded steel cabinet from an old Meissner Signal Shifter. When completely buttoned-up the whole thing is almost a solid cube. I use relay rack panels for the 1/8-inch material and work it with a hacksaw, electric drill, file and ordinary chassis punches.

Another mechanical weakness in many *commercial and homebuilt receivers* is the tuning mechanism. The homebuilder can use one of the rugged military surplus mechanisms or pick up an old National PW dial and drive as I did for the DBR-1. A tuning mechanism with a solid, smooth feel gives the impression of quality much as the "thunk" of a well-fitted door does to an automobile.

The plug-in coil assemblies are from an old National HRO receiver. They are excellent for experimenting with different coils. You can try the effect of different coil form materials on oscillator drift, for instance, or determine the effect of different antenna couplings on front-end selectivity. The coil drawers were sawed and chiseled from an old HRO chassis.

mixer circuit

The ideal mixer should be able to *simultaneously* handle the strongest and weakest signal you will encounter while contributing little noise itself and providing sufficient conversion gain to reduce the noise of the i-f strip to insignifi-

cance. Not many mixers meet these requirements. The 7360 does to a greater degree than most and is used in the DBR-1.

The overall noise factor of the mixer plus i-f strip is

$$F = F1 + \frac{F2 - 1}{G1} \qquad (1)$$

where F is the overall noise factor of the complete receiver, F1 is the noise factor of the mixer, F2 is the noise factor of the i-f strip and G1 is the power gain of the mixer.

Assume the 7360 mixer with its preselector circuits has an 8-dB NF (6.3X) and the i-f strip has a 13-dB NF (20X), including the 6-dB insertion loss of the input filter, and the mixer power gain is 10 dB (10X).

$$F = 6.3 + \frac{20 - 1}{10} = 8.2X \text{ or } 9.2 \text{ dB}$$

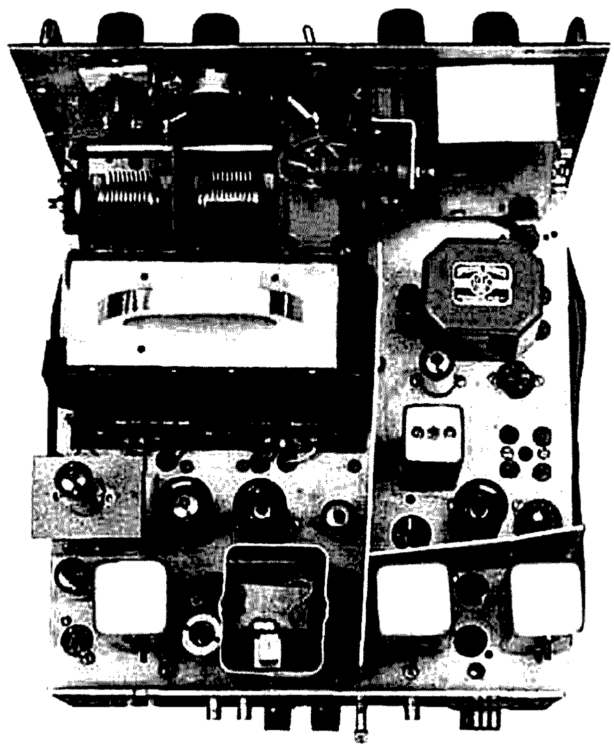
Suppose you contemplated using a diode balanced mixer instead of the 7360 into the same i-f strip. Suppose also that the diode mixer has the same 8-dB NF as the 7360 but it has no gain (G = 1). Therefore,

$$F = 6.3 + \frac{20 - 1}{1} = 25.3X \text{ or } 13.9 \text{ dB}$$

A screwdriver adjustable pot on the front panel is for experimenting with the effect of bias on the sensitivity and overload characteristics of the 7360. In my location two 50-kW locals, WBZ on 1030 kHz and WHDH on 850 kHz, produce a strong third-order IM product on 1210 kHz which is convenient for evaluating the linearity of the mixer. At one bias point, which is quite sharp and may vary from one 7360 to another, the IM product nulls out. On the tubes with which I have experimented the point of maximum immunity to IM is also very close to the point for maximum sensitivity.

vfo circuit

The vfo and buffer amplifier use two sections of a 6U8A. A 20-pF varactor diode connected across the oscillator tank circuit provides a vernier tuning control to precisely set the carrier. The control voltage comes from the regulated +105-volt line and is further regulated with a 12-volt zener diode.



Top view of the DBR-1 shows the 1/8" chassis plate which is bolted to the front panel and the rear apron. The i-f strip runs along the back with filter FL2 in the open shield can in the center. From left to right below the coil drawer are V8, V1, V2, V7, V6 and Q1. Below the output transformer are the voltage regulator, V9, and CR3 and CR4.

The National tuning capacitors which are an integral part of the PW gear drive have straight-line frequency characteristics. When the DBR-1 is used as a tunable i-f with the 3.5- to 4.0-MHz coils, the calibration is 1kHz per division of the PW dial. The 0.9- to 2.0-MHz coils give 2 kHz per division across most of the dial.

i-f strip and filters

The i-f strip provides the required adjacent channel selectivity with enough gain to drive the detector on the weakest signal and should have a low enough noise figure to prevent degrading the NF of the

mixer. The noise factor (ratio) of a properly designed i-f strip is

$$F_{i\text{-}f} = F1 + F2 \qquad (2)$$

where $F_{i\text{-}f}$ is the overall noise factor of the i-f strip, F1 is the insertion loss of the input filter and F2 is the noise factor of the first i-f tube.

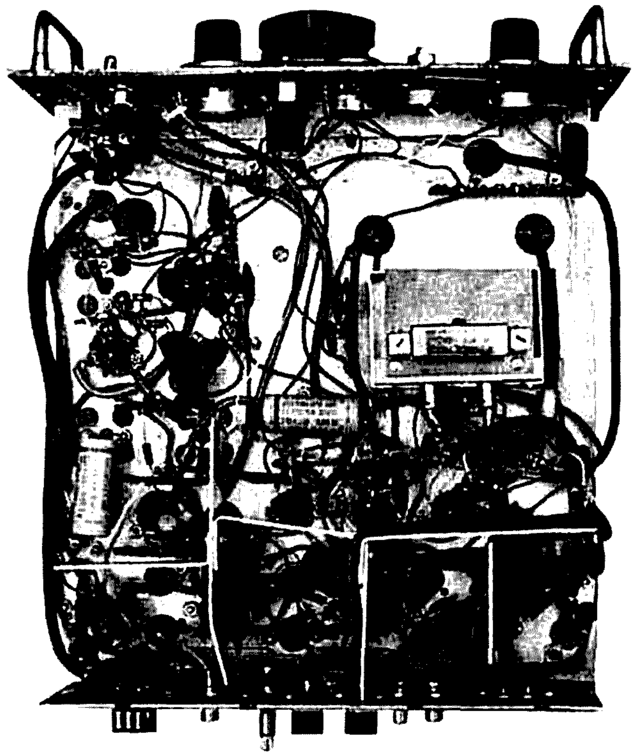
Cascading mechanical or crystal lattice filters allows you to achieve near perfect shape factors and stop bands down more than 120 dB. The National Radio Club has done considerable work on the cascading of mechanical filters[5] and the club periodically purchases batches of Kokusai filters which it matches and sells to members in sets of two or three. Each Kokusai filter comes with a complete set of characteristics and the current units are extremely good.

Kokusai filters are used in the DBR-1. Filter FL1 has a center frequency of 454.98 kHz, a bandwidth of 2.58 kHz at the 6-dB points and a −60 dB bandwidth of 4.50 kHz, resulting in a 6/60-dB shape factor of 1.74:1. Filter FL2, at the same points, is 455.00 kHz, 2.48 kHz, 4.26 kHz and 1.72:1.

The combined characteristics of the two filters in the DBR-1, checked both by the point-by-point method and by the sweep method to be described later, are 2.1 kHz at −6 dB, 2.45 kHz at −10 dB and 2.8 kHz at −60 dB, resulting in a 6/60 dB-shape factor of 1.33:1 and a 10/60 dB shape factor of 1.14:1. The 10/60-dB shape factor is given because it seems to describe more accurately the *useful* nose bandwidth and gives a better idea of the *steepness* of the skirts than does the usual 6/60-dB figure.

For several reasons it is best to distribute the filters in the i-f strip rather than connecting them back-to-back.[6,7] One of the filters should be at the input to the i-f strip to satisfy the requirement of putting the adjacent channel selectivity as close to the antenna as possible. Additional filters should be placed one or two stages further along.

If all the filters were placed at the input to the i-f strip the following amplifiers could generate considerable wideband noise outside the bandpass of the filters. Also, two or more cascaded filters are capable of stop bands more than 120 dB down and if they are lumped together it requires a lot of isolation between the input and output. If they are separated the isolation requirements can be reduced to 60 dB at a step. Another consideration is that two or three filters at the input to the i-f strip could degrade the NF of the strip enough to degrade the NF of the entire receiver.



Bottom view of the DBR-1 receiver. Filter FL1 is in right center. Note the shielding around each i-f stage at the bottom. Rf chokes at bottom left filter the incoming B+ and bias voltages.

A good bet for a filter for a high frequency i-f is the Swan SS-16. This 16-pole crystal-lattice filter is centered on 5.5 MHz. The 6-dB bandwidth is 2.7 kHz and it has a 6/60-dB shape factor of 1:28:1 and a 6/120-dB shape factor of 1/8:1. The stop band is 140-dB down.

Stop bands of −120 and −140 dB sound great and they are not too difficult to achieve. In practice, however, anything over 100 or 120 dB is of little practical

use because there are mighty few front-ends which will simultaneously handle two signals 120-dB apart (in strength). For instance, 140 dB above 1 μV is 10 volts and most receivers will block completely long before the signal gets up to 10 volts. Thus, the fact that the filter could separate the 1-μV signal from the 10-volt signal is only of academic interest because it will never get the chance.

The filters must be complemented by a meticulous job of shielding, isolating, bypassing and filtering to prevent degradation of the filter characteristics. In the DBR-1 each stage is individually shielded and double decoupled and the heater leads are bypassed. A copper shield is soldered between the grid and plate pins of each i-f tube socket. These measurer insure a quiet and electrically stable i-f strip that takes full advantage of the filter characteristics.

checking the i-f passband

The most accurate and simple method of evaluating an i-f strip is to plot its passband and skirts point-by-point with a signal generator, counter and attenuator. This is fine for getting a picture of the final results, but it isn't something you want to repeat continuously during the alignment process.

The requirements for sweeping a steep-sided, narrow passband are much more stringent than those for sweeping a tv i-f strip. You must sweep at a very slow rate, 1 Hz or less, to show the steep sides and sharp corners. Commercial firms use special sets of equipment made by people like Rhode & Schwartz that sell for $10,000 or more.

Fortunately for home-builders there is a $130 piece of equipment that, with a simple variable frequency signal source, will do almost the same job. This is the Heath SB-620 Spectrum Analyzer. My experience is limited to the SB-620 connected for a 455-kHz i-f, but it should give similar results at higher frequencies.

Connect the end of the i-f strip to the i-f *input* jack on the rear apron of the SB-620 through a 5-pF, or less, capacitor. Place the function switch on the rear apron in the *ham scan* position. Set the *variable sweep rate* and the *variable sweep width* controls at the maximum counterclockwise positions and the *sweep width* switch in the 50-kHz position. At these settings the sweep rate is about 2 Hz and the horizontal calibration of the screen is about 0.7 kHz per division.

Connect a variable frequency signal source to the input of the i-f strip and manually sweep it *slowly* (say 5 seconds per sweep) back and forth across the passband. The peak of the pip will trace out the passband of the i-f passband on the face of the CRT (see fig. 2).

Another method which is useful for examining the skirts of the passband is to feed the output of a noise generator into the i-f strip with the SB-620 connected and adjusted as before except that the *sweep width* switch should be in the 10-kHz position. The result is a trace of the passband as shown in fig. **3**. The top of the trace is jagged due to the random amplitude of the noise pulses but you get a good picture of the skirts which can be extended down to −60 dB by using the *−20 dB log* position of the *amplitude scale* switch.

table 1. Comparison between some of the performance characteristics of a typical amateur-band receiver and the better military and professional class receivers.

	i-f feedthru	image ratio	stability	IM	i-f stop band	front-end dynamic range	frequency read-out
commercial amateur receiver	50 dB	60 dB	100 Hz per hour	60 dB	60 dB	60 dB	1 kHz
best professional & military receivers	100 dB	100 dB	1 Hz per month	100 dB	100 dB	130 dB	1 Hz

Optimum reception of weak a-m signals in the presence of noise, QRN and QRM demands the exalted-carrier technique. In addition, the exalted carrier must be phase locked to the original carrier so that the technique of chopping off the a-m carrier with the filter in a ssb receiver and then using the bfo to replace it is ineffective. You must either lock a locally generated carrier to the original carrier or filter, process and amplify the original carrier.

In the DBR-1 a 15- or 20-dB peak is superimposed on one edge of the passband with a Q-multiplier.[8] The carrier of the desired station is placed on the peak, exalting it in relation to the sidebands. The series resistor between the Q-multiplier and the i-f transformer which is advocated in the referenced article is not used because the resistor restricts the height of the peak to only about 10 dB which is not sufficient.

When used for carrier exaltation the Q-multiplier is best placed at the very end of the i-f strip. Here it is protected from off-frequency stations by the filters and it "sees" all inband signals at about the same strength since the gain of the receiver is largely controlled by the i-f stages. The Q-multiplier is double decoupled and completely shielded and has parasitic supression resistors in both the gate and drain leads. Voltage is taken through a divider from the +105 volt line. The result is a remarkably stable Q-multiplier which can be set on the verge of oscillation, if desired, and remain there indefinitely.

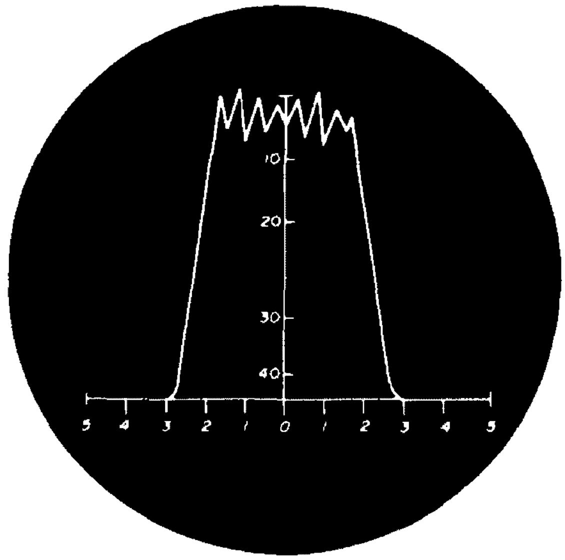


fig. 3. I-f passband of DBR-1 as displayed on SB-620 with white noise fed into i-f strip input. This gives good picture of the skirts which can be extended down to -60 dB with the built-in 20 dB attenuator. Horizontal calibration is about 0.5 kHz per division.

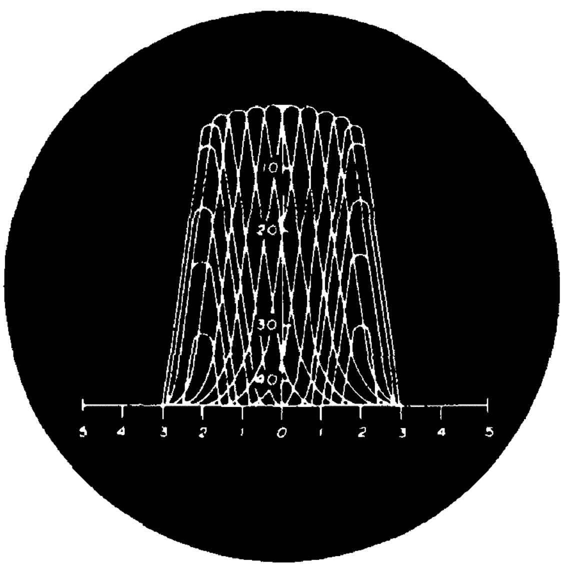


fig. 2. I-f passband of DBR-1 as traced on *SB-620 by slowly* sweeping a CW signal *across* the i-f. Note that the sides of the passband are defined by the ascending and descending row of peaks at each side, not by the skirts of the pips. In practice only one pip is seen at a time, first climbing up one side, then across the top, and finally down the other side. Horizontal calibration is about 0.7 kHz per division, vertical in dB.

The Q-multiplier is fix tuned to either the upper or lower edge of the passband and can be switched from one to the other for sideband selection. A small trimmer is mounted on the front panel for those occasions when the receiver is used as a tunable i-f. The Q-multiplier is then made to oscillate to serve as a bfo for ssb reception and the trimmer is used to position the carrier down the side of the passband.

audio amplification

This section of the receiver is often taken for granted, yet proper attention to

detail here can make a significant improvement in the readability of signals. It has been shown that if the high speech frequencies are cut it is necessary to also cut some of the lows to maintain optimum readability.[9,10] In a properly designed ssb transmitter the lows are cut in the audio and filter stages. A-m broadcasts, on the other hand, carry the full audio range out to at least 5000 Hz. If you cut the highs at 2500 Hz and leave the lows, the speech will sound boomy and muffled. Roll off the lows below 300 or 400 Hz and the speech becomes crisp and readable.

Various exotic audio filters were tried in the DBR-1 with expensive high-Q inductors but the simple trick of using a small coupling capacitor to the audio output stage worked as well as any in practice. The switch marked *audio filter* shorts out the capacitor if the full response of the amplifier is desired.

The audio noise limiter shown in the diagram is a solid-state copy of the limiter used in the Collins R390. Like all other audio limiters and clippers tried in the DBR-1 it is ineffective because of the filters and Q-multiplier ahead of it. Experiments with i-f limiting just before the second filter have been much more favorable. The idea is to limit all signal and interference components in the passband to the level of the desired sidebands up to the point of carrier exaltation.

conclusion

The DBR-1 is a tailor-made receiver for my interests and location. It is superior to any receiver available commercially for the purpose for which it was designed and in the environment in which it is used. It is superior only because the market for such a receiver is so limited that no manufacturer could afford to design and build a receiver for that purpose. The future of homebuilt receivers is in this area of "performance" or "competition" — receivers which are designed for ultimate operating characteristics for a specific, limited task.

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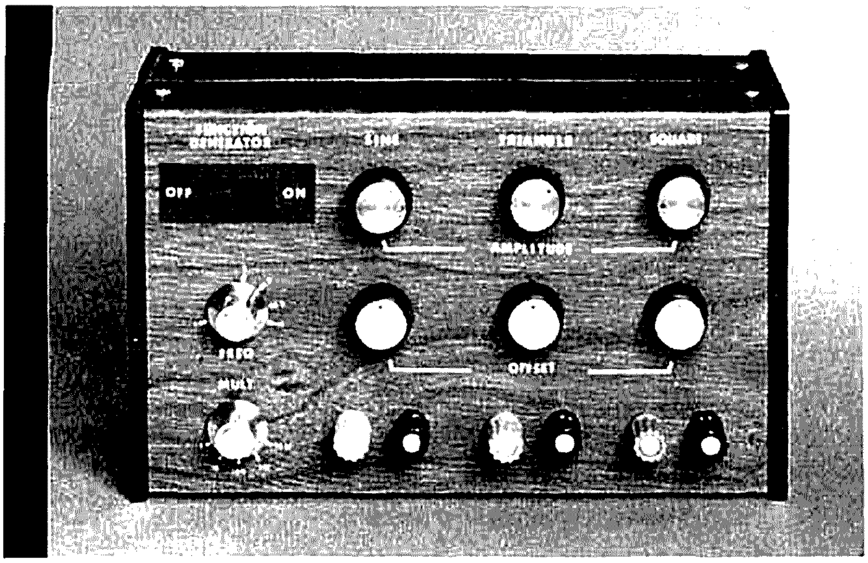
ham radio



integrated-circuit function generator

Construction of a variable function generator that provides sine, square and triangular waveforms over the frequency range from 1 Hz to 100 kHz

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

I recently had need for a second audio generator when I became involved with some circuits requiring two different and simultaneous audio frequency inputs. After considering the alternatives, I decided this would be as good a time as any to put to use a new waveform generator IC that I purchased some time ago. I expected that the IC would minimize the time required to go from drawing to finished generator. This proved to be the case. I highly recommend the function generator described here to anyone who needs a compact audio signal generator.

general description

The heart of this instrument is a waveform generator IC manufactured by Intersil Inc. and known as the 8038.* The IC will produce sine, square and triangle waveforms over the frequency range from 0.001 Hz to 1.0 MHz. Frequency may be continuously varied with a pot, programmed with fixed resistors or swept with a voltage ramp. The function generator described here covers the range from 1.0

*The version of the 8038 IC used here is the 8038CC which is an economy copy selling for under $4.00 in singles. Since Intersil products are not widely distributed through outlets available to most amateurs, drop me a line and let me know if you wish to purchase one or more of these ICs. If the demand is sufficient I will order enough to take care of requests. An sase would be appreciated.

Hz to 100 kHz in 5 decade bands. A pot is used as the vernier element and a 5-position switch as the band selector.

Since the outputs from the *8038 IC* are not all of uniform amplitude nor at useful power levels, an amplifier was added to each output along with trimmers for adjusting the output to a standard level. The amplifiers each consist of an op-amp and booster which is capable of driving low-impedance loads.

One final feature which was included is adjustable offset for each of the outputs. This function may or may not be useful to you, but I find it helpful on occasion and decided to include it since it only required the addition of 3 pots and 3 resistors.

8038 IC

This IC contains quite a bit of circuitry with over 50 transistors, some diodes and dozens of resistors squeezed onto the chip to accomplish the desired functions. Fig. 1 is a block diagram of the IC showing the relationships between the major sections of the waveform generator.

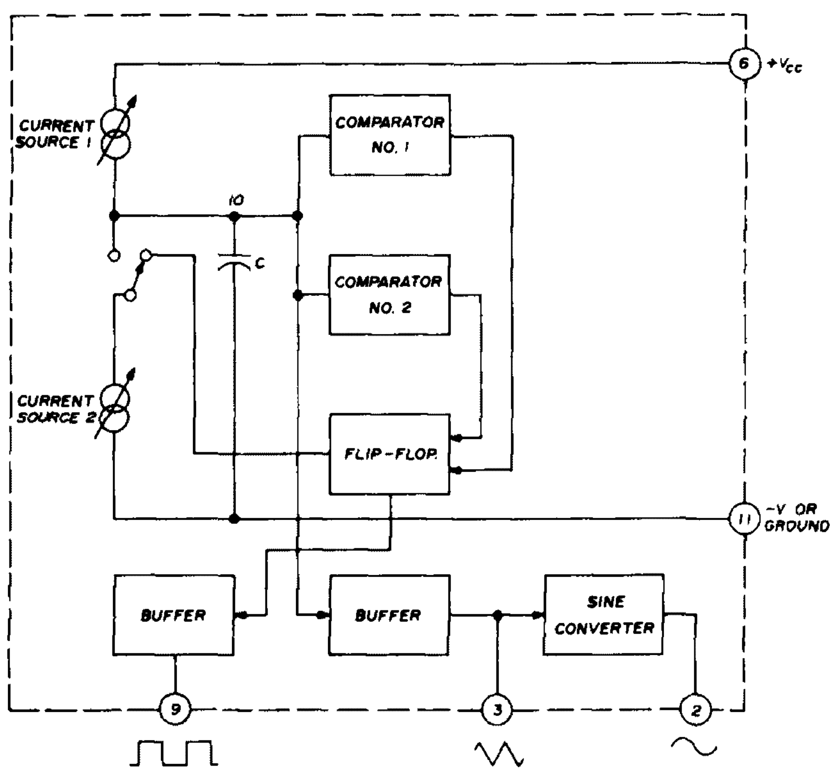


fig. 1. Block diagram of the 8038 waveform generator.

Initially, current source 2 is disconnected and an external timing capacitor, C, is charged through current source 1. When the voltage across C reaches the trigger threshold of comparator 1, an output is generated which causes the flip-flop to change state and current source 2 becomes active.

As the capacitor discharges towards a negative peak, it reaches the firing threshold of comparator 2 whose output resets the flip-flop, disconnecting current source 2 and permitting the cycle to repeat. The current level supplied by the two current sources is controlled by either a series resistor to the power supply or an external bias voltage.

The bias voltage controls both sources simultaneously and is normally used for fm or swept operation of the oscillator. Since the power supply connections for the current sources terminate at separate pins, current levels may be controlled independently with separate resistors. This latter method is used when unsymmetrical waveforms such as a pulse or sawtooth is desired.

By varying the ratio or charge to discharge time of the timing capacitor, the square-wave duty cycle may be varied from 2% to 98% and the triangular wave adjusted for either a positive or negative sawtooth or ramp.

The *waveform appearing across the* capacitor is internally fed to a buffer amplifier and then to both the sine converter and the output (pin 3). With a triangle input to the sine converter, total harmonic distortion of the resulting sine wave is typically less than 1%. With proper adjustment distortion levels as low

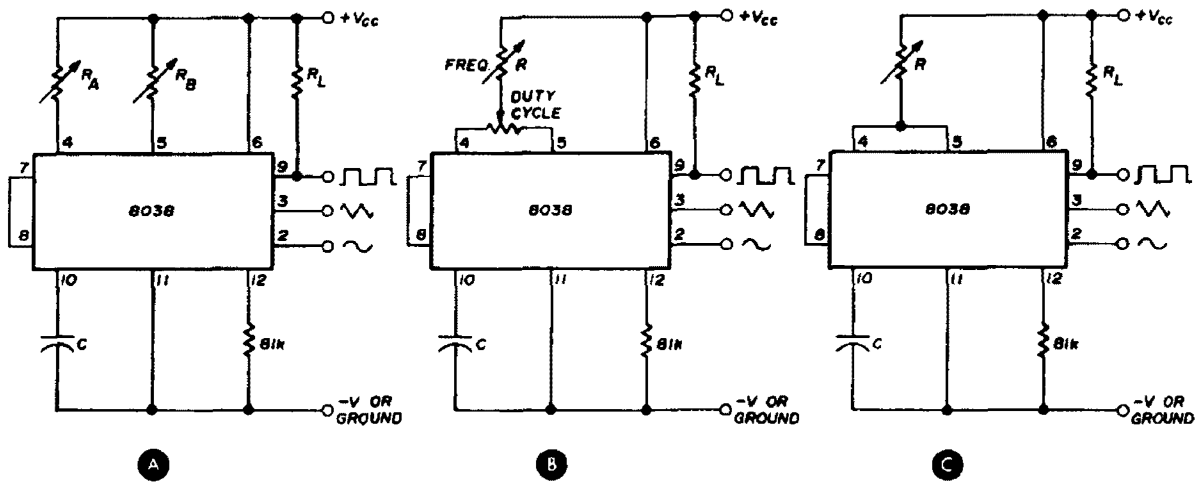


fig. 2. Three possible methods for connecting external timing resistors.

as 0.5% are possible, according to the data sheet.

The square wave is taken from one side of the flip-flop and is fed to a buffer amplifier having an uncommitted collector. Supply voltage for this stage is thus independent of the rest of the circuitry and allows use of a separate 5-volt supply for TTL compatibility, if desired.

The IC will operate from either a single supply of from 10 to 30 volts or dual supplies of ±5 to ±15 volts. With dual supplies all outputs will be symmetrical about ground.

external connections

Fig. 2 shows three possible connections for the external timing resistors at pins 4 and 5. Potentiometer R_A controls the rising portion of the sine and triangle and the zero state of the square wave. The method used in fig. 2A is most desirable for large changes in duty cycle while that shown in 2B is used mainly to control frequency with minor changes in duty cycle. A 50% duty cycle results when $R_A = R_B$ and if no adjustment is contemplated, the connection shown in fig. 2C may be used. Resistance values for R_A and R_B may vary from a minimum of 500 ohms to a maximum of 1.0 megohm.

Both frequency modulation and sweeping may be applied to the waveform generator. Fig. 3 illustrates how this is accomplished. For small deviations the modulating signal is applied to pin 8 through a coupling capacitor. Pin 7 is an internally-set bias voltage which is normally applied to the current sources by a direct connection between pins 7 and 8.

The input impedance at pin 8 is around 8000 ohms and may be raised by inserting a resistor between the two pins as shown. For large deviations, or sweep-

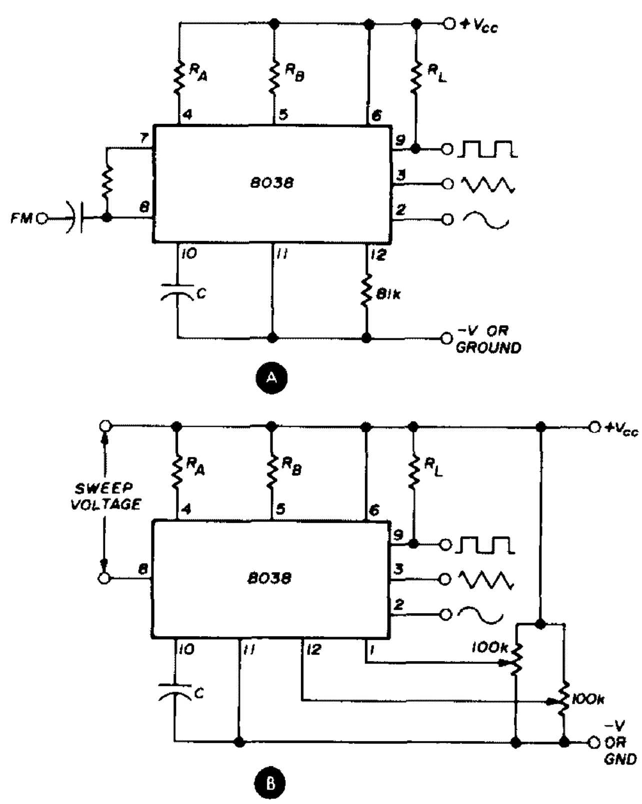


fig. 3. Connections for fm (A) and sweep operation (B) of the 8038 IC.

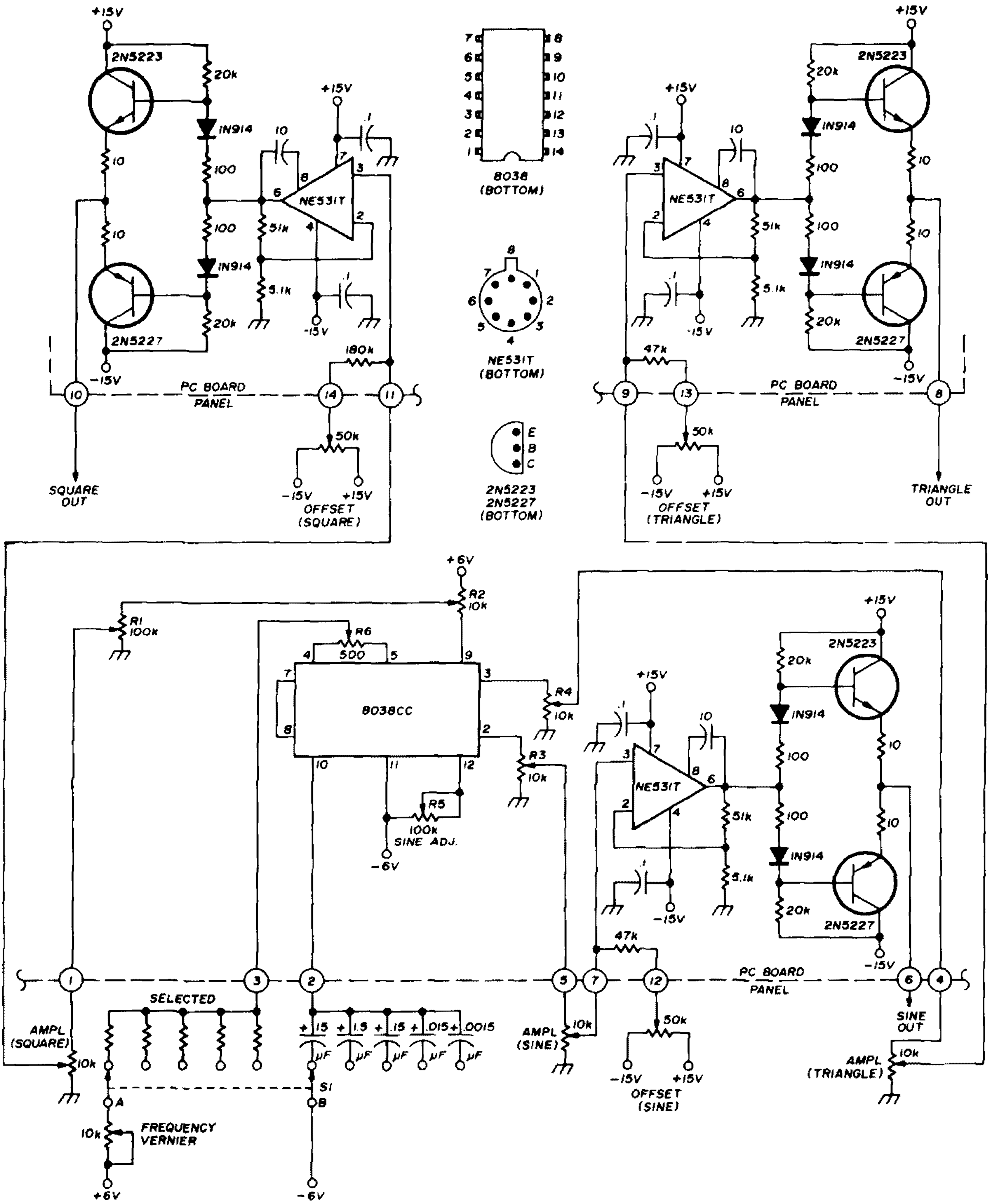


fig. 4. Circuit diagram for the function generator. Numbered leads correspond to holes provided in the PC board. All resistors are ¼ watt, 5%.

ing the oscillator, pin 7 is disconnected and the sweep voltage is applied between the positive supply and pin 8 as in fig. 3B. The sweep voltage must be confined within the limits V_{cc} and $2/3\ V_{cc}$.

It may be noted that in most of the diagrams discussed so far, an 81k resistor is shown connected from pin 12 to the negative supply. Pin 12 is a sine wave adjustment point and 81k is an approxi-

mate value for minimizing distortion. If a 100k pot is substituted for this resistor, a more accurate adjustment is possible. Additional improvement in correcting sine wave distortion is possible by applying correction at both pin 1 and pin 12 as illustrated in fig. 3B.

in the output afford some protection but it is not recommended that the output be short-circuited on a regular basis.

Neither fm nor sweeping operation were incorporated in this instrument since I had no need for these functions. My other audio generator has wide range

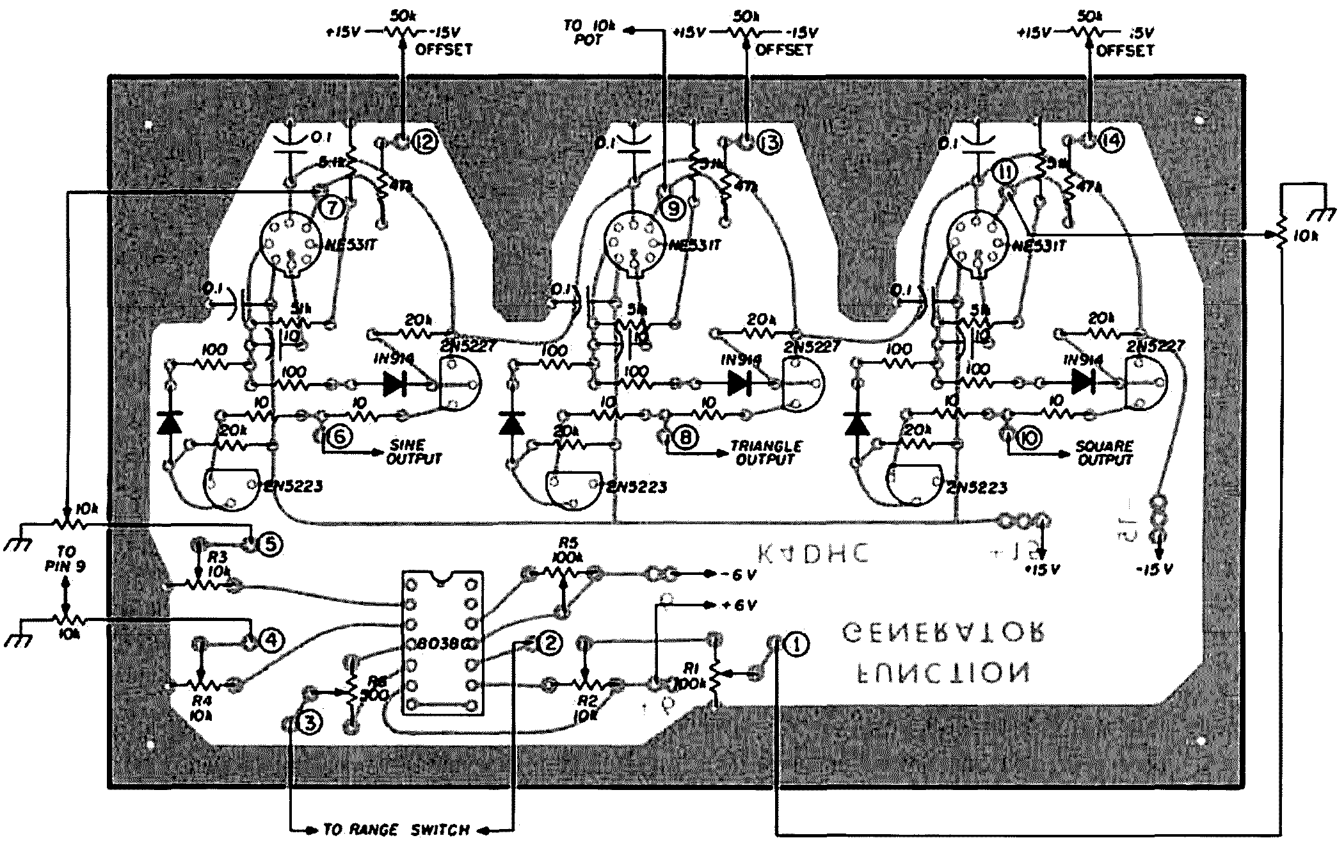


fig. 6. Component layout for the function generator board. Numbered holes correspond to numbers on the schematic for external connections. All diodes are 1N914 or equivalent. Printed-circuit is shown in fig. 8.

function generator

The circuit for the function generator is shown in fig. 4. To achieve the desired bandwidth of 100 kHz, a high performance op-amp was required for the three output stages. The Signetics NE531T was chosen since it is available from several sources at a reasonable price. This amplifier requires only a single external capacitor for compensation and has a full-power bandwidth well beyond 100 kHz. It also shows no tendency towards instability.

The current boosters are a pair of complementary transistors with diode/resistor networks in the input to prevent crossover distortion. The 10-ohm resistors

sweep capability as well as CW operation so the function generator was kept to CW only. Of course, there is no reason why these functions could not be included if required and the preceding discussion should be helpful towards that end.

Power supplies need not be regulated since frequency is not dependent on voltage as far as the IC is concerned. If external bias is applied to pin 8 for either sweeping or setting frequency, however, stability will be affected by any fluctuations in this control voltage. It should be regulated.

Frequency drift of the vco with temperature is 50 ppm/°C typical. This is

quite good and applies over the range from 0°C to 70°C. In order to maintain low dissipation and allow little heating of the 8038, the IC is run at the relatively low voltage of ±6 volts. The remaining circuitry operates at the full ±15 volts.

The circuit for the power supply is shown in fig. 5. A PC-type power transformer was used since it was available, but a suitable substitute could be mounted in the same space if not too large. The power supply circuit board layout and parts placement are shown in fig. 7.

construction

All leads from the front panel of the instrument are routed directly to holes provided on the printed-circuit board. The five timing capacitors and five calibrating resistors are mounted right on the rotary switch assembly. The resistors are not installed until calibration has been completed, at which time the proper values will have been determined. The printed-circuit board is 5.6-inches (14.2 cm) long by 3.5-inches (8.9 cm) wide. A full-size layout is shown in fig. 6.

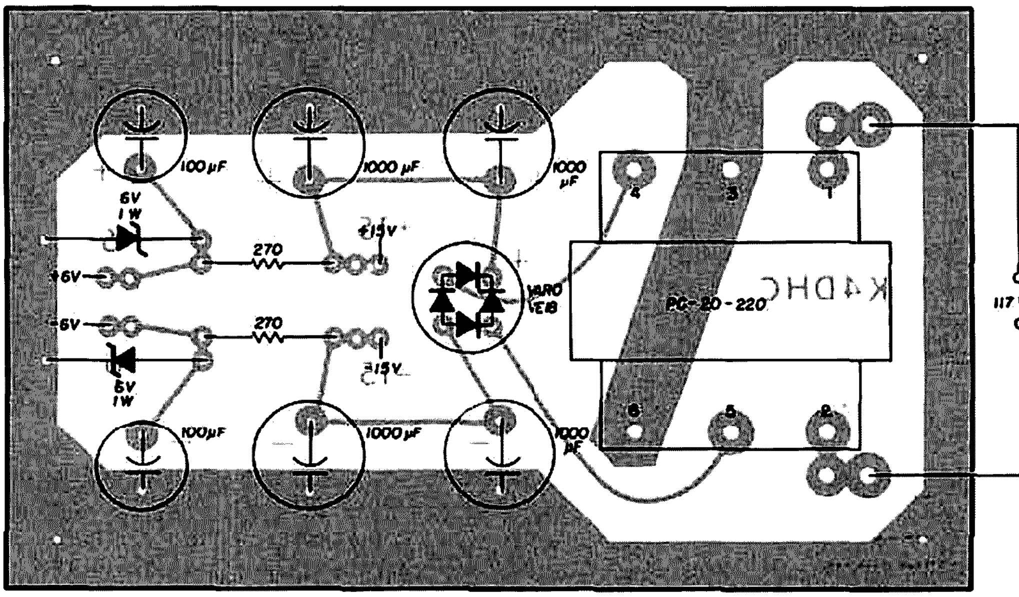


fig. 7. Full-size component layout for the power supply. Printed circuit is shown in fig. 8.

calibration and adjustment

When you are ready to fire up the function generator, preset all the trimmers on the PC assembly to mid-position and temporarily connect a 1k pot in series with the arm of S1A and lead no. 3 from the PC board. Set the pot for maximum resistance and turn the band switch to position 3 (100 to 1000 Hz). Turn the frequency control pot to the high end and the offset controls, if used, to mid-position (zero offset).

The triangle will be checked first since it will most clearly indicate proper symmetry. Connect your scope to the triangle output and apply power to the

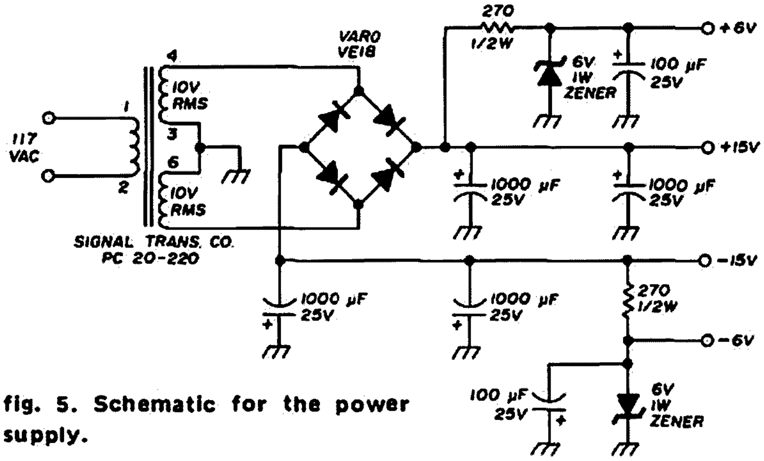


fig. 5. Schematic for the power supply.

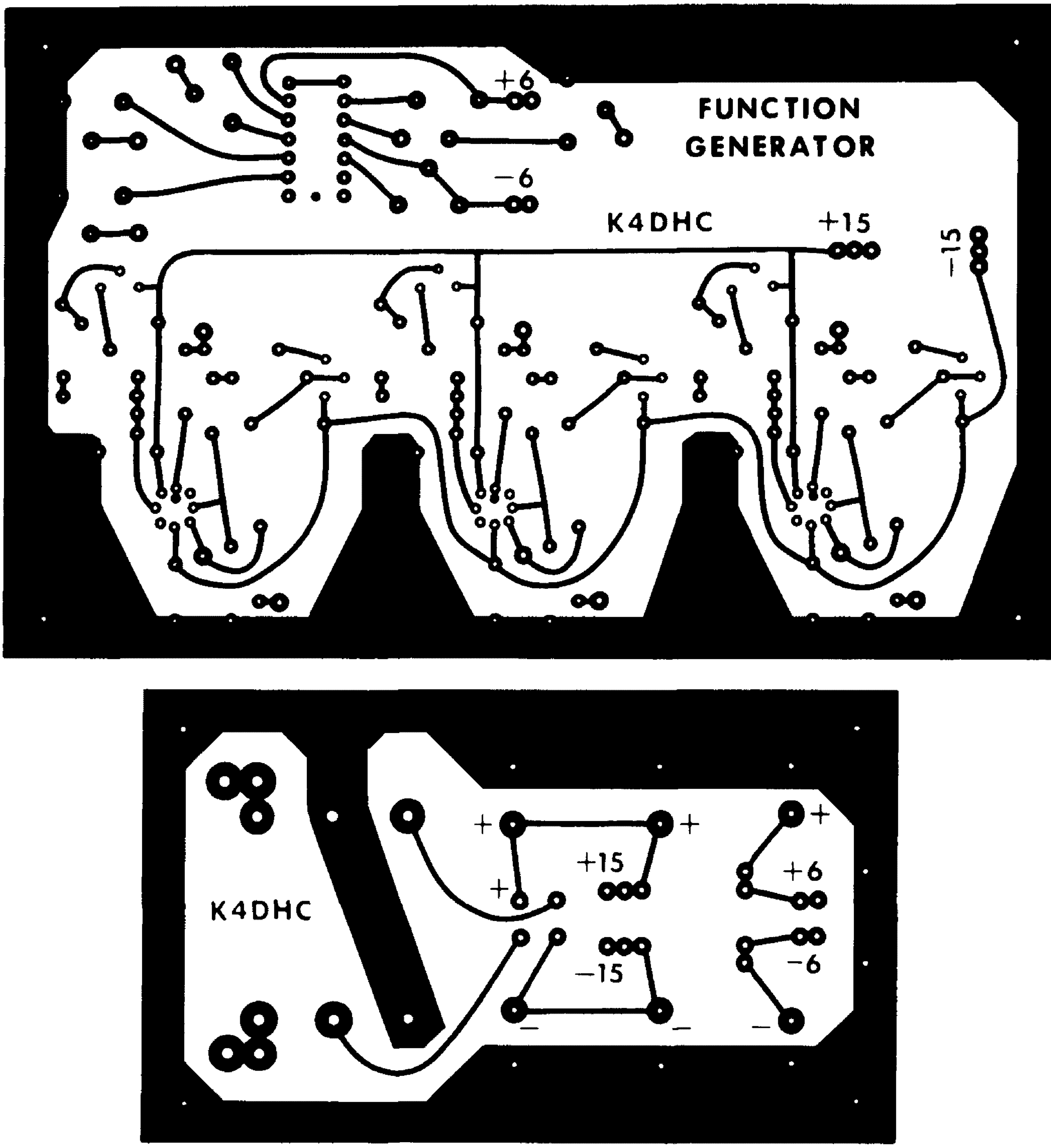


fig. 8. Full size printed-circuit layouts for the function generator (above) and power supply (below).

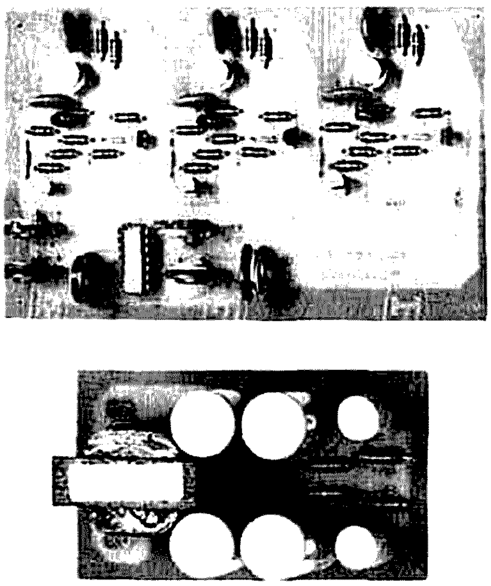
generator. With the amplitude control at maximum, adjust trimmer R4 for a 10-volt peak-to-peak output. Adjust R6 for proper symmetry and then the temporary 1k pot for a frequency of 1.0 kHz. If a slight overlap is desired, adjust for a slightly higher frequency. Oscillator frequency with the vernier control at minimum should be 100 Hz or less.

The sine wave should be adjusted next, so transfer the scope probe to the sine wave output. With the amplitude at maximum, adjust R3 for a 10-volt peak-to-peak output. Adjust trimmer R5 for minimum distortion of the sine wave.

Next look at the square wave output. Both R2 and R1 must be adjusted for the desired 10-volt peak-to-peak output signal. Trimmer R2 will cancel out the offset voltage present at the square wave

output and R1 will control the amplitude. There is interaction between these two controls, but juggling back and forth a few times will accomplish the desired results.

Measure the resistance of the temporary pot and make a note of the required value for this band. Go to each of the remaining bands in turn and



The completed function generator and power supply circuit boards.

determine a value of resistance to calibrate each one. Remove the 1k pot and install fixed calibrating resistors at the proper points on the bandswitch.

The 1.5- and 15-μF timing capacitors for the two lowest frequency bands should preferably be tantalum types. The remaining capacitors may be mylar or polystyrene. If any band gives calibration trouble, it may be due to a capacitor with a wide variation from the marked capacitance value and a replacement may be necessary. The trimmer resistors are the blue plastic vertical PC mounting type made by CTS and available from many distributors or at Radio Shack stores. The 0.1-μF bypass capacitors are 50-volt discs while the 10-pF units are small dipped micas.

ham radio

the need for coherent frequency-shift keying in RTTY

A discussion of modern communications theory, and how it may be used to improve the performance of RTTY communications

Stephen A. Maas, K3WJQ, 1195B Bear Mt. Drive, Boulder, Colorado 80303

In recent years many modern designs have been presented for high performance RTTY demodulators.[1,2,3] These units have used a-m demodulation techniques, acknowledged to be superior to fm, and incorporate excellent autostart features. Unfortunately, many RTTY enthusiasts consider these demodulators to be optimum for demodulating signals in noisy environments not realizing that, with slightly more elaborate signal-processing techniques, it is possible to achieve considerably better performance, especially when the signal-to-noise ratio (snr) is low. One such technique, as yet little explored in amateur radio communications, is the use of coherent demodulation.

snr improvement

It is universally recognized that there are severe limits in improving the snr at the receiver input — the most common method is to increase transmitter power. Those amateurs who admire the trend toward low power and solid-state transmitters are not at all pleased with this approach; it is expensive, limited to one kilowatt, and worst of all, does nothing to improve reception at your own station.

With an rf preamplifier using modern transistors, the noise figure of virtually any amateur communications receiver can

be reduced. However, in all but the most mediocre receivers electronic noise is insignificant as compared to atmospheric noise and radiation from the sky and earth. Also, it would be possible to improve RTTY performance by increasing the duration of the mark and space pulses because the probability of a receiver error (mistaking a mark pulse for a space pulse or vice versa) decreases with increase in pulse energy. Because of the standardization of mark and space pulses, however, this approach is impossible. This leaves signal processing as a technique for improving RTTY performance. The use of coherent techniques over envelope demodulation of the RTTY mark and space pulses offers the best practical improvement.

RTTY demodulation

Frequency shift keying (FSK) or radio-teletype (RTTY) is a type of modulation where the transmitted signal alternates between two distinct frequencies. Hence, it may be treated as a digital form of fm and demodulated by passing the signal through a limiter and discriminator. Alternatively, FSK may be considered as two on-off keyed signals and can be demodulated with a-m techniques; that is, using no limiter and an envelope or product detector.

Because of their limiter, fm systems are capable of practically eliminating noise when the input snr is good. However, fm-based systems exhibit a threshold effect, and below a certain critical snr the signal becomes completely lost in noise. For this reason a-m systems, which at worst exhibit only a minute threshold effect, are preferred when difficult noise and interference conditions exist (the usual conditions on most amateur bands).

The block diagram of a high-performance a-m FSK demodulator is shown in fig. 1. The first stage is a filter, used to minimize interference, which admits both mark and space signals. It is followed by separate mark and space filters and envelope detectors; these are diode detectors followed by simple low-pass filters, and their output is the envelope of the input pulse (see fig. 2). The output of one channel is negative with respect to ground, however, and the other positive. These outputs are then added and presented to the threshold corrector and slicer, which together form a sort of "decision" apparatus.

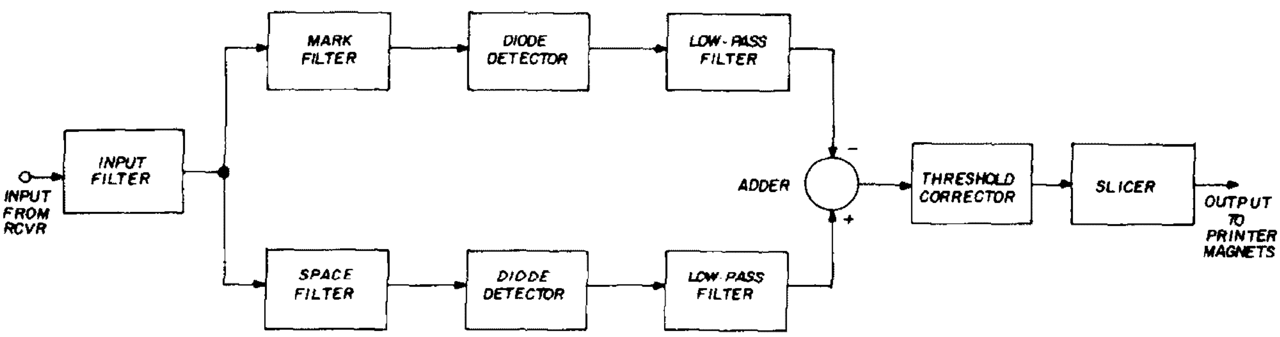


fig. 1. Block diagram of an RTTY demodulator based on a-m techniques.

The purpose of the decision stages is to decide which type of pulse, mark or space, was received. When the snr is good, there is little problem. However, if the snr is low, there is great uncertainty as to whether a mark or space pulse was received. One of the larger random fluctuations in the noise could cause a mark pulse to be mistaken for a space pulse, or vice versa, and the printer would print an improper letter.

In making this decision, you are faced

with a problem. Unlike the signal, the noise voltage cannot be given a deterministic mathematical representation; however, its probability distribution *is* known (the probability that the noise voltage will not exceed any given value at a given time). During the interval in which you expect to receive a pulse, you observe a certain voltage at the output of the adder. Knowing the probability distribution of the adder output voltage (which can be calculated, although with difficulty), you can determine whether the voltage is more probably caused by a mark or space, and decide accordingly.

If the voltage is above a certain value, a space was probably sent; if below, a mark. This middle value, where the probabilities are equal, is the threshold voltage* and is determined by the threshold corrector.

For those readers with a mathematical turn of mind, the detector output voltage, when a signal is applied, has a Ricean probability density. Noise alone has a Rayleigh density. At any time, one channel has signal plus noise, the other has noise only, so the output of the adder will have the convolution of the Rice and Rayleigh densities as its probability density function.

If no independent fading of the mark and space signals occurred, both channels were of equal gain, and both signals of equal amplitude, the proper threshold voltage would be zero; that is, choose *mark* if the adder output is negative, and choose *space* if positive. There would be no need for the threshold corrector. However, the mark and space channels do undergo independent fading, so a non-zero threshold value is computed. The usual value of threshold voltage is chosen to be halfway between the average value of mark channel output voltage and the average space channel output voltage:

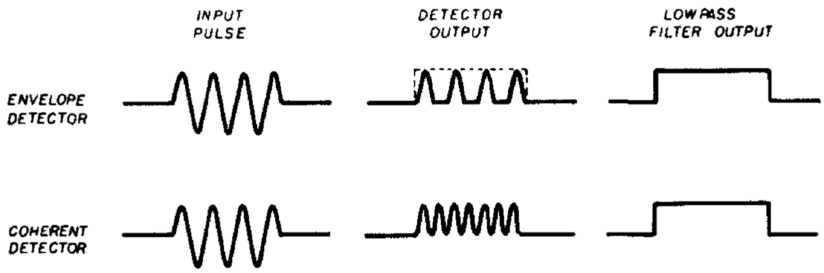


fig. 2. Operating characteristics of envelope detectors and coherent detectors.

table 1. Percent correct copy vs input signal-to-noise ratio.

	probability of error		percent correct copy	
input snr	coherent FSK	non-coherent FSK	coherent FSK	non-coherent FSK
0 dB	0.15	0.30	32%	8%
2 dB	0.10	0.25	48%	13%
3 dB	0.05	0.15	70%	32%
6 dB	0.01	0.10	93%	48%
9 dB	10^{-5}	0.01	99.9%	93%

$$V_{thresh} = \frac{\overline{V_{mark}} + \overline{V_{space}}}{2}$$

Remember, V_{mark} is negative and V_{space} is positive. The line above V_{mark} and V_{space} indicates an average value of voltage. This threshold voltage is at best a close approximation of the optimum, which is a complicated relation, practically impossible to implement electronically.

coherent demodulation

An FSK demodulator using coherent

*Not to be confused with threshold effects mentioned in connection with fm; these are entirely different phenomena.

detection is shown in **fig. 3**. The post-detection filters are identical to those used in the non-coherent demodulator. Instead of using an envelope detector, the received signal is multiplied in a product detector by a sinusoid of the same frequency *and* phase. Thus, the phase information in the signal, which is ignored by the envelope detector, is used in the coherent detector to improve the signal-to-noise ratio at the detector output.

The improvement over non-coherent FSK varies from 4 dB at low signal-to-noise ratios to 2 dB at high (greater than 6 dB) snr. This means that using coherent detection instead of noncoherent detection results in an improvement equivalent to an extra 4 dB of antenna gain or transmitter power. In terms of percent correct copy there is a noticeable improvement (see **table 1**). It is clear that coherent modulation results in readable copy in situations where the best non-coherent FSK systems would print gibberish.

An added advantage, not considered in calculating **table 1**, is that

$$V_{thresh} = \frac{V_{mark} + V_{space}}{2}$$

is precisely the optimum threshold for a coherent system. This arises from the fact that the probability density for the adder output voltage is a Gaussian function (bell-shaped curve), unlike that of the non-coherent system.

The disadvantage of coherent demodulation is that a signal identical in phase to the received signal must be generated. This task is best accomplished by a phase-locked loop. A detailed discussion of phase-locked loop operation is beyond the scope of this article (and beyond the scope of most communications theory textbooks, for that matter) but a simple description may be presented. You may consult the references for further information.

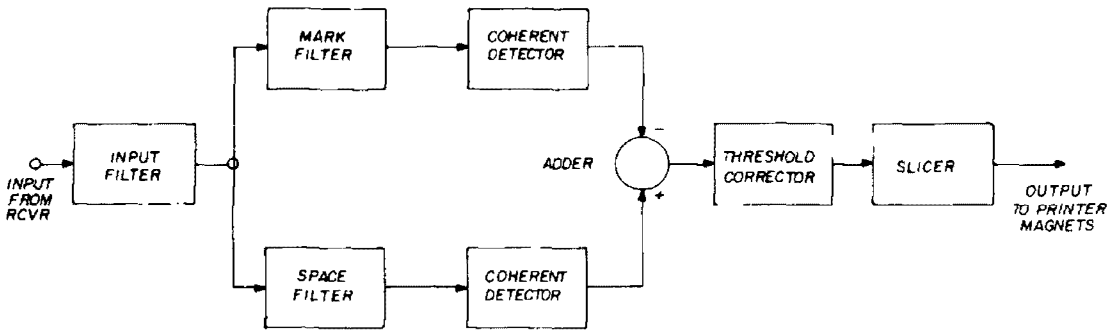


fig. 3. Block diagram of a coherent a-m RTTY demodulator.

The phase-locked loop consists of a multiplier (product detector), lowpass filter, and voltage-controlled oscillator (vco) (see **fig. 5**). The vco output and the input signal are applied to the product detector; its output is a beat frequency or, if the vco frequency and signal frequency are the same but the phases differ by other than 90°, a dc voltage. This voltage is applied to the vco which changes frequen-

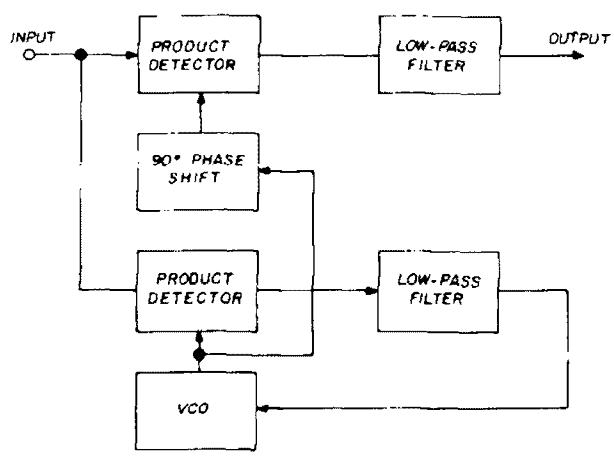


fig. 4. Coherent detector using a phase-locked loop.

cy by an amount proportional to the applied voltage until the signal and vco frequencies are the same and their phases are close to 90° apart. The result is that the vco output follows phase changes of the applied signal. The vco output, shifted in phase by 90°, is used for the

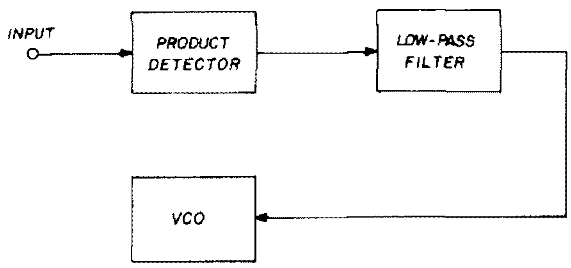


fig. 5. Basic phase-locked loop consists of a vco, product detector and lowpass filter.

heterodyning signal in the coherent detector.

The coherent detector requires complicated circuitry. However, high quality phase-locked loops ICs are available commercially, some with capabilities for a-m as well as fm demodulation. Hence, the entire detector and phase-locked loop can be had with a single integrated circuit. One such IC, available on the surplus market for about $3.60, is the Signetics NE561. Thus, for a total investment of about $7.00 over the cost of a conventional a-m demodulator, you can obtain the best possible RTTY demodulation.

transmitter modifications

In general, no transmitter modifications should be necessary. It is possible to build a phase-locked loop demodulator which will lock onto the mark or space pulse well within its 22-millisecond duration. However, during the time the loop is not locked, signal-to-noise ratio at the detector output is not optimum. This results in lowered performance.

To minimize locking time, the phase of the mark and space pulses should be as constant as possible. Therefore, instead of one oscillator switched between two tuned circuits, the transmitter should use two free-running oscillators, one to generate mark signals, the other for space signals. Keying is accomplished by switching between the outputs of the two oscillators. In this way the phase-locked loops will not have to re-lock onto each successive pulse. It should be emphasized, however, that the coherent system described here will work with conventional FSK transmitters with superior performance as compared to that of non-coherent reception.

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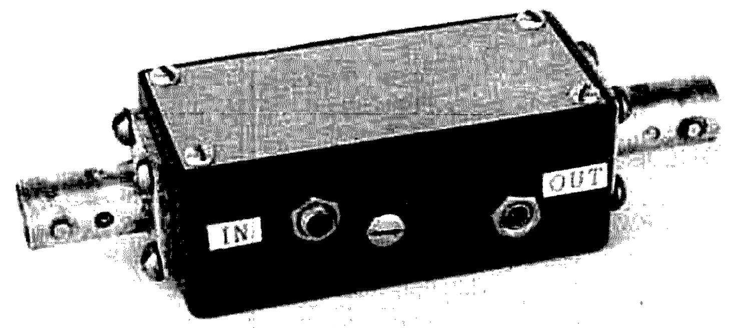
ham radio

"I see it's almost time for you and your ham set to renew your marriage license."



a good two-meter preamplifier

Construction of a high-performance 144-MHz preamp that provides 15-db gain and low noise figure

Ronald E. Guentzler, W8BBB, Route 1, Box 30, Ada, Ohio 45810

Recently I had need for a small, mechanically-stable, well-shielded, low-noise, two-meter preamplifier. The first three requirements were met by using a Pomona 2397 "Black Box." It is a cast-aluminium box measuring 2.2 x 2.8 x 5.7 cm (7/8 x 1-1/8 x 2-1/4 inches). Because of the small size, direct wiring rather than a printed circuit was used; this might partially account for the good noise figure obtained.

There are three important contributors to the noise figure (NF) of a preamp.

1. The gain of the preamp and the NF of the stage following it.[1] This preamp has 15-dB gain which is adequate so long as the unit it is feeding is reasonably good; too much gain between the antenna and the first mixer is bad.[2]

2. The transistor. The RCA 3N159 has a respectable NF rating.[3] It is selected by the manufacturer to have a maximum NF of 3.5 dB and a typical NF of 2.5 dB at 200 MHz; the performance will be better at 147 MHz. The 3N200 performs as well as the 3N159.

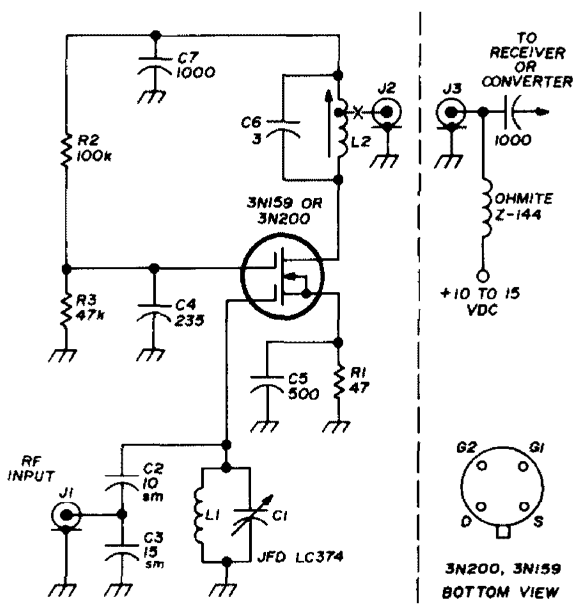
3. The input circuit. The best transistor is worthless unless a low-loss (and thus low-noise) input circuit is used. The best input circuit coupling arrangement is the "tapped capacitor" and the worst is link coupling.[2] A good coil must be used. This circuit uses the JFD LC374 which is a combination piston capacitor and an *air supported, silver plated* coil.

the circuit

The preamp circuit is shown in fig. 1

and partial parts layouts are given in **fig. 2**. Capacitors C2 and C3 provide the input coupling and present the proper "turns ratio" for impedance transformation as well as providing most of the capacitance for resonating the input circuit; both should be silver micas. L1 and C1 are the JFD LC374. C1 is used mainly to physically support L1 and one transistor lead, but it also provides a means for adjusting the resonant frequency of the input circuit. If all is proper, the input circuit should resonate with C1 near to its minimum capacitance.

A dual-gate mosfet was used because it does not require neutralization (so long as adequate shielding, etc., are provided); this eliminated one coil and a lot of trouble. The 3N159 is the lowest noise RCA unprotected dual-gate mosfet but since it does *not* have internal gate protection, caution should be observed in handling the device. Of the protected RCA dual-gate mosfets, the 3N200 has the best NF specification. However, the 3N200 has higher input and output capacitances than the 3N159. Therefore, if a 3N200 is used in this circuit, use an 8-pF silver-mica for C2 and a 12-pF silver-mica for C3.*



C1 part of JFD LC374 tank circuit (see text)
C2 10-pF silver-mica with 3N159, 8-pF silver mica with 3N200
C3 15-pF silver-mica with 3N159, 12-pF silver mica with 3N200
C4 235-pF mica button
C5 500-pF mica button
L1 JFD LC374 tank circuit (contains C1)
L2 6 turns no. 22 enamelled on a 5-mm (0.2") diameter slug-tuned form, tap at 1 turn

fig. 1. The two meter preamp. Power is fed into the preamp through the rf output connector. On the right side is shown a power feed arrangement to be located inside the unit (receiver, converter, etc.) with which the preamp will be used.

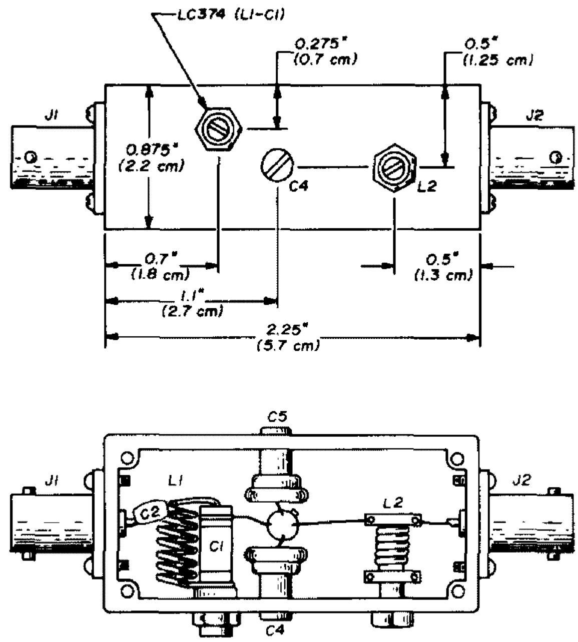


fig. 2. Permanently mounted component location. As viewed from above, the coil on the JFD LC374 is almost beneath its capacitor, C1. The transistor is located between the mica button capacitors. Make sure that the transistor will fit between the capacitors before mounting them.

construction

As can be seen in **fig. 2**, C4 and C5 are button-mica capacitors. They support two of the transistor leads. Their values may seem small, but these values were chosen to provide series resonance with the transistor leads.

The value of R1 was optimized to give minimum NF with 10-volts dc applied to the preamp. However, if other considera-

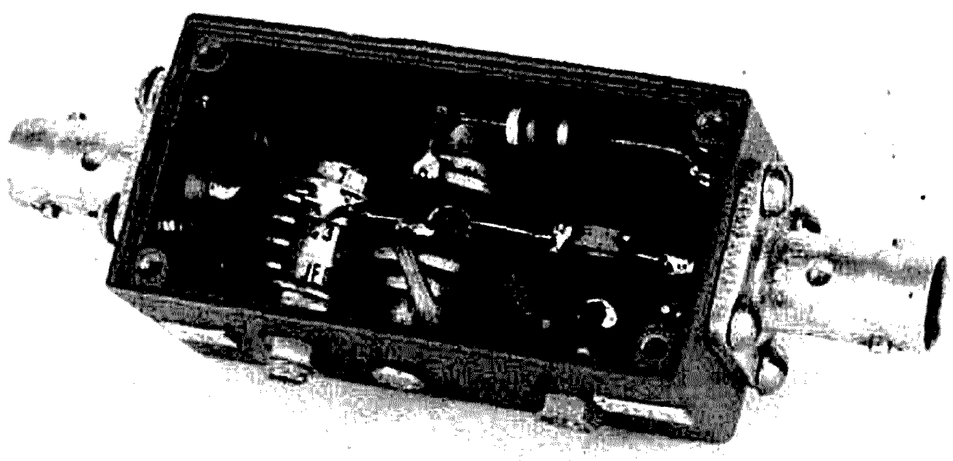
*The Texas Instruments 3N204 and 3N211 appear to be good alternates for the RCA 3N159 and 3N200. Although the 3N204 has a lower gain than the 3N211, it has a better noise figure. The author has not yet tried the TI transistors in the preamp. Editor.

tions are important, it can be increased in value.[3]

The output tank is a conventional slug-tuned coil, 5-mm (0.2-inch) diameter, with 6 turns number-22 enamelled wire, tapped 1 turn from the cold end. Depending upon the coil form and slug characteristics, C6 might have to be changed to obtain the proper resonant frequency.

alignment

Alignment is quite simple. Adjust C1 and L2 for maximum signal. Then adjust C1 for minimum noise figure. (Note that tuning can be accomplished from the outside with the cover on.) If noise measuring equipment is not available, the approximate minimum noise figure can be obtained by tuning C1 on the low frequency side of the resonance (run the piston into C1); tune either for approximately a 10% decrease in output signal or turn the piston into C1 by about one-half turn.



Construction of the low-noise two-meter preamp showing the component layout. Case is a small cast-aluminum box made by Pomona (see fig. 2).

It was considered undesirable to provide a separate input for the dc feed because of the small physical size of the preamp. Therefore, dc power is fed through the output coax (this is desirable for antenna-mounted preamps). An arrangement for feeding the dc into the coax is shown on the right side of fig. 1. If direct dc feed is desired, C7 can be replaced by a feedthrough capacitor such as a Centralab FT-2300; then, a 1000-pF ceramic should be inserted in series with the output connector at the point indicated by an X.

To assure good contact between the coax connectors and the box, the threads in the connector flanges were drilled out, and the box was drilled and tapped for 4-40 machine screws. The screws projecting through the box provided a good means for obtaining ground points.

The good results were obtained by using the Pomona "Black Box" (actually, it is blue), the JFD tank, the dual-gate *mosfet*, and the tapped-capacitor input circuit. Don't cheat yourself—buy the proper components.

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ham radio

optimum height for horizontal antennas

How to choose antenna height to put your signal where you want it to go

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

A recent magazine article[1] showed that the higher the better holds true for the gain of a horizontal dipole, and presented a graph showing improvements of over 20 dB with increased height. Since 20 dB is a pretty healthy gain, I decided a further consideration of this approach to increasing antenna gain was called for.

Any antenna book will show that the distant field strength from a horizontal dipole is proportional to sin (h sinα), where h is the electrical length and α is the radiation angle. This equation is, however, too simplified for many practical antenna comparisons. As is often the case, I fall back to Kraus' Antennas[2] and his equation (11-87) on page 305 is pertinent here. This equation gives the gain for a half-wave horizontal antenna above ground referenced to a half-wave antenna in free space. Thus it allows comparison to be made to a standard reference, as well as comparison of one real antenna to another.

The equation is:

$$G = \sqrt{\frac{R_{11} + R_{1L}}{R_{11} + R_{1L} - R_m}}\ |2 \sin (h_r \sin \alpha)|$$

R_{11} = $\lambda/2$ self resistance

R_{1L} = $\lambda/2$ loss resistance (assumed zero here and in Kraus' *Antennas*)

R_m = mutual resistance between $\lambda/2$ and its image 2h away

h_r = $(2\pi/\lambda)h$

Where R_{11} is 73 ohms at an infinite height, and $(R_{11} - R_m)$ is the antenna resistance at any given height, as shown in

fig. 2-45 of the ARRL Antenna Handbook,[3] or Kraus, page 305.

determining factors

This equation shows that the difference in gain between two antennas of different height occurs at certain combinations of height and radiation angle, that the antenna resistance is an important term when the usual assumption of constant power is used, that electrical height should be used, and that the maximum gain of a real antenna over one in free space is 2 to 1.

How, then, can you get more than 6 dB gain when the maximum gain over a free-space antenna is only 2 to 1? It is a case of comparing apples to oranges. The 2 to 1 improvement figure refers to the comparison of the field strength of a certain height real-life horizontal dipole at the radiation angle giving maximum field strength to a similar one in free-space. The over-6 dB case involves comparing the field of one real antenna to the field of another real antenna, at a certain radiation angle. We will see, however, that a 20-dB improvement does not always result, depending upon what combination of a, $(R_{11} - R_m)$, and h_r is used.

table 1. Vertical-radiation from various height horizontal antennas, as compared to a half-wave dipole in free space (see fig. 1). The last column shows angle at which maximum radiation occurs.

antenna height (radians) h_r	(wavelength) $\frac{h}{\lambda}$	vertical radiation angle a 2°	a 10°	a 20°	a 30°	a 60°	a 90°	a max
36°	0.1	0.08	0.41	0.79	1.15	1.93	2.19	90°
90°	0.25	0.10	0.49	0.94	1.29	1.79	1.83	90°
108°	0.30	0.11	0.56	1.05	1.41	1.74	1.65	56.4°
126°	0.35	0.13	0.64	1.17	1.53	1.63	1.39	45.6°
180°	0.50	0.22	1.06	1.79	2.04	0.83	0	30.0°
270°	0.75	0.32	1.45	1.98	1.40	1.60	1.98	19.5°
360°	1.00	0.44	1.77	1.67	0	1.49	0.0	14.5°

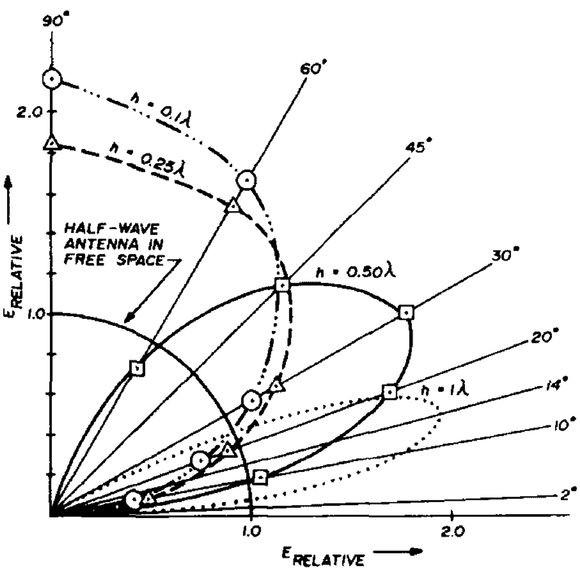

fig. 1. Relative field strength for horizontal half-wave antennas vs h in wavelengths as a function of radiation angle for constant power.

The equation shows that at any height between 0.1 and 0.25 wavelength the maximum field of a horizontal is always at a radiation angle of 90° (straight up!), although the low-angle radiation does increase at the 0.25 wave height. Above a quarter wave each height, h, has an angle where the maximum radiation occurs, as shown below:

h/λ	a
0.25	90°
0.30	56.4°
0.35	45.6°
0.50	30.0°
0.75	19.5°
1.00	14.5°

This shows why it is so hard to get good low-angle radiation with a horizontal dipole on 80 meters. This is also what got me interested in using verticals for

80-meter DX. Antenna height in wavelengths should be used with the formula, or at least the height in feet should be converted to wavelengths, as the results will then apply to any amateur band. A 10-foot (3-meter) high dipole works a lot differently at 80 meters than one at that height for the 20-meter band.

The equation, or better yet the principles upon which it is based, shows zero radiation at a radiation angle of zero degrees. This is because the antenna and image have currents equal in strength, but opposite in phase, so the field at a distance cancels at $a = 0^\circ$. There is, however, radiation at angles for $\alpha > 0^\circ$. This radiation is plotted in fig. 1.

examples

To illustrate these points, consider some examples using an h of 0.1 wavelength as compared to one of 1 wavelength, at several radiation angles. The field from either is zero at $a = 0^\circ$. Next use $a = 3^\circ$.

The gain of an antenna 0.1 wavelength high compared to one in free-space at $a = 3^\circ$ is:

$$G = \sqrt{\frac{73}{21}}\ |2 \sin (36^\circ \times \sin 3^\circ)|$$
$$= 1.86 \times 2 \times 0.0329 = 0.1226$$

For the 1-wavelength high antenna at $a = 3^\circ$,

$$G = \sqrt{\frac{73}{73}}\ |2 \sin (360 \times \sin 3^\circ)|$$
$$= 1 \times 2 \times 0.3229 = 0.6459$$

The gain of the 1-wavelength high antenna over the 0.1-wavelength height is (0.6459/0.1226) = 5.27, or 14.4 dB. This is about the best improvement that I calculated at this radiation angle. Other 10 to 1 height ratios gave the following gains:

0.5/0.05	10 dB
0.35/0.035	3 dB
0.6/0.06	11 dB

It is possible that there could be 20-dB improvement at smaller radiation angles, but the actual field-strength values would be quite small at such low angles. Let's try $a = 14.5^\circ$, where the 1-wavelength antenna height has radiation maximum.

$$G \text{ for } h = 1\lambda \text{ and } a = 14.5^\circ \text{ is } 2.0$$
$$G \text{ for } h = 0.1\lambda \text{ and } a = 14.5^\circ \text{ is } 0.5828$$
$$G\ (1\lambda/0.1\lambda) = 10.7 \text{ dB}$$

At 30°, the 1 wavelength high antenna has no radiation, so the 0.1 wavelength height would be a better choice and 0.5 wavelength would be even better since its maximum radiation occurs at $a = 30^\circ$. Any number of examples could be worked out, but all of these appear in fig. 1. Using it or the data in table 1, you should be able to select the optimum antenna height for the radiation angle of greatest interest to you.

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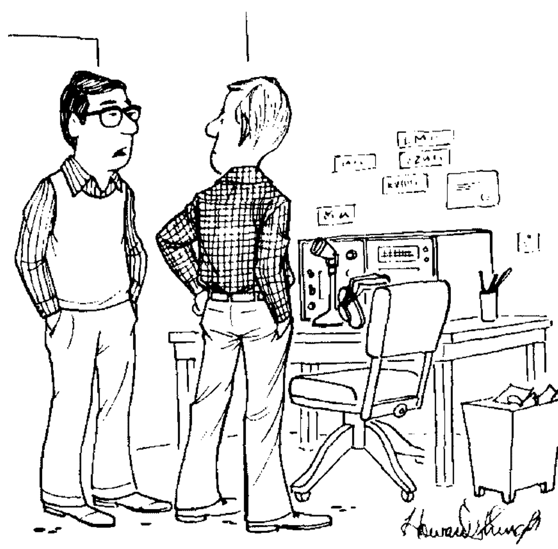
ham radio



"I'm expecting to experience strong interference on an unwanted sideband . . . My mother-in-law is coming to visit."

local-oscillator waveform

effects on spurious mixer responses

Tests with doubly-balanced IC mixers and dual-gate mosfet mixers indicate no detrimental affects with square-wave injection signals

Max Robinson, K4ODS, John W. Smith, Route 8, Box 105, Bowling Green, Kentucky 42101

Up until now receiver local oscillators have been crystal or LC controlled and the waveform was of no particular significance. The output of these oscillators was a fairly good sine wave. Everyone accepted, without question, that the local oscillator output should be *clean.* Current receiver designs require that the first local oscillator be crystal controlled at 500 kHz intervals throughout the high-frequency spectrum. The frequency synthesizer seems to be a natural for this but it can introduce some problems.

The output of a synthesizer is often a square wave, and in the usual tunable i-f receiver, the use of a tuned circuit to clean up the waveform could become very complicated. The question has arisen in my mind, can I get away with using a square-wave local oscillator?

I feel, as do most hams, that one good test circuit is worth a thousand words of theoretical discussion. I decided to try two different mixer circuits. The first was an integrated circuit doubly-balanced mixer, the Signetics N5596, which is equivalent to the Motorola MC1596.[1] The other circuit I tried was a dual-gate mosfet, the MPF122. The reason for using these circuits was that I planned to

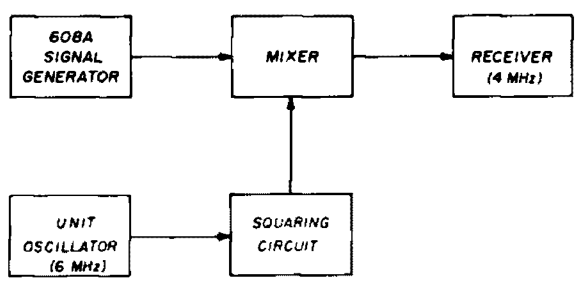


fig. 1. Test setup for comparing performance of N5596 IC mixer and dual-gate mosfet mixer with square-wave local oscillators.

use them in some projects which are now in the planning stage.

test procedure

The source of sine-wave local oscillator was a homebrew signal generator built around a General Radio Unit Oscillator (fig. 1). The square-wave source was derived from the sine-wave source using the circuit shown in fig. 2. The 10-turn pot is a symmetry adjustment and in all cases was adjusted for minimum second harmonic response of the mixer circuit under test. The signal source was a Hewlett-Packard 608A signal generator which has a minimum frequency of 10 MHz. The receiver used was a Drake 2B.

In all tests, the local oscillator was set to 6 MHz and the receiver was set to 4 MHz. The signal generator was first set at 10 MHz to obtain a beat with the fundamental of the local oscillator, and the signal level was adjusted for a reading of 20 dB over S9 at the receiver. Next, the 608A signal generator was set to obtain a beat with the harmonics of the local oscillator injection frequency and the 608A. The attenuator was reset to maintain the former S-meter reading at the receiver.

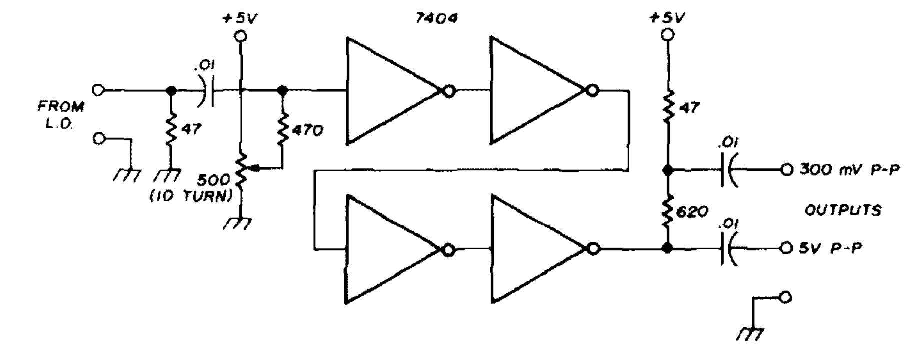


fig. 2. Simple IC circuit for generating a square-wave output from the sine-wave generator.

For example, the third-harmonic spurious response of the mixer was checked by setting the 608A at 14 MHz to beat with the third harmonic of the injection frequency. Then the signal level was increased until the S-meter once again read 20 dB over S9. If the signal level required was at or above the overload level, the data point was thrown out.

The N5596 IC was connected as

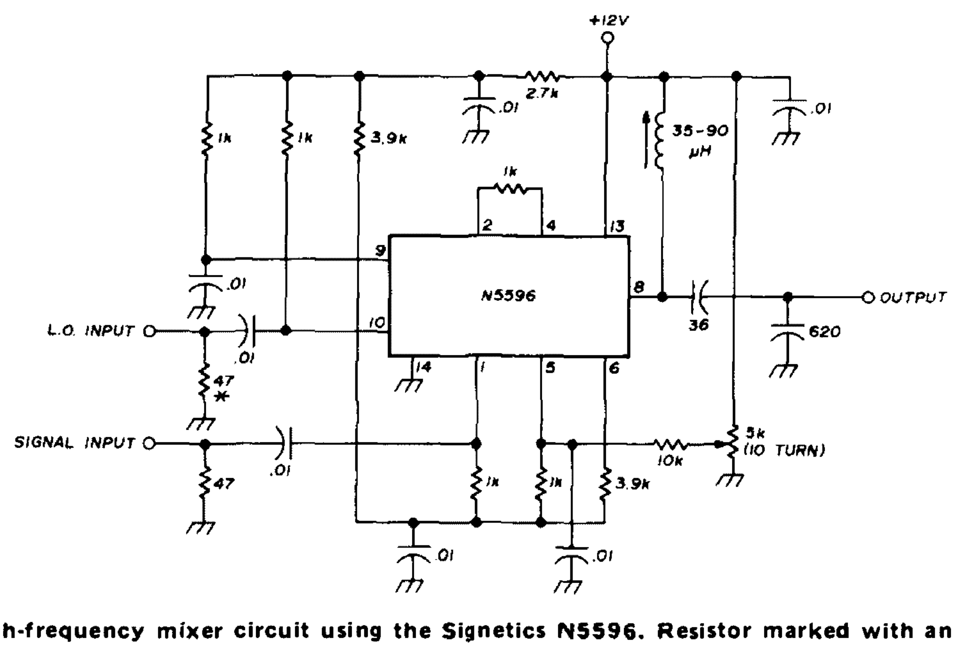


fig. 3. High-frequency mixer circuit using the Signetics N5596. Resistor marked with an asterisk is removed when using a square-wave local-oscillator.

shown in fig. 3. The balance control was adjusted by setting the local oscillator to 4 MHz and adjusting for minimum carrier feedthrough. Table 1 shows the results of the tests. The spurious responses are given in dB relative to the fundamental response at 10 MHz. The injection voltages given in the column headings are according to data supplied by Signetics.[2]

In the table, the notation *OL* means that an overload signal level was required to produce a 20 dB over S9 meter reading. The note *NM* means that no measurement was taken at this frequency.

The dual-gate mosfet mixer is shown in fig. 4. I did not have very much data on the MPF122 so I guessed that the output impedance would be about the same as the N5596. The value of the source resistor comes from a page of the information supplied by Circuit Specialists Company. Gate two was placed at dc ground because this gave maximum conversion gain.

The chosen injection voltage of 1-volt rms seemed to be optimum. The 5-volt p-p square wave was chosen to maintain the same sine-wave voltage to square-wave voltage ratio as that used for the N5596 IC.

As a final check, the square wave was viewed on an oscilloscope and it was confirmed that the minimum second harmonic response really did occur when the square-wave was symmetrical. The results of testing the MPF122 mixer are given in table 2.

table 1. Harmonic response of N5596 IC balanced mixer in dB below fundamental response.

harmonic number	frequency (MHz)	dB with 60 mV rms sine-wave injection	dB with 300 mV p-p square-wave injection
1	2	NM	NM
1	10	0	0
2	8	NM	NM
2	16	-40	-44
3	14	-19	- 9
3	22	-17	- 9
4	20	OL	-41
4	28	OL	-48
5	26	-34	-12
5	34	-33	-10
7	38	OL	-14
9	50	OL	-14
11	62	OL	-13
13	74	OL	-12
15	86	OL	- 7

conclusions

One of the more obvious conclusions is that the local oscillator should be operated above the incoming signal. For the frequencies used here, the received frequency would be 2 MHz. The closest spurious signal is at 8 MHz which could be attenuated by a preselector of good design. On the other hand, if the 10-MHz signal were used, the 8-MHz spurious could be rather troublesome.

In both mixer circuits, using a square-wave injection actually reduces the second harmonic product. If the preselector has very good ultimate attenuation, the use of square-wave could be advantageous.

The harmonics of very high order are much stronger with a square wave than with a sine wave and although this may not cause any trouble in a receiver, the presence of vhf spurious signals in a transceiver or transmitter could make TVI reduction more difficult. If a premixer and fixed first i-f are used, the first mixer injection waveform can be easily cleaned up by simply adding another gang to the preselector.

A type-D flip-flop might be a useful premixer because its output is only the

table 2. Harmonic response of dual-gate MPF122 mosfet mixer in dB below fundamental response.

harmonic number	frequency (MHz)	dB with 1-V rms sine-wave injection in	dB with 5-V p-p square-wave injection
1	2	NM	NM
1	10	0	0
2	8	NM	NM
2	16	-15	-47
3	14	-12	-10
3	22	-13	- 9
4	20	-23	-41
4	28	-22	-43
5	26	-23	-13
5	34	-22	-12
7	38	NM	-18
9	50	NM	-15
11	62	NM	-21
13	74	NM	-15
15	86	NM	-10

difference frequency (plus odd harmonics) with no sum or original frequencies coming through.

afterthoughts

The article by Moore[3] arrived too late to be factored into the preparation of this article. He points out that one manufacturer of professional receivers uses square-wave injection voltage to the mixer. This adds relevance to the work done in this article.

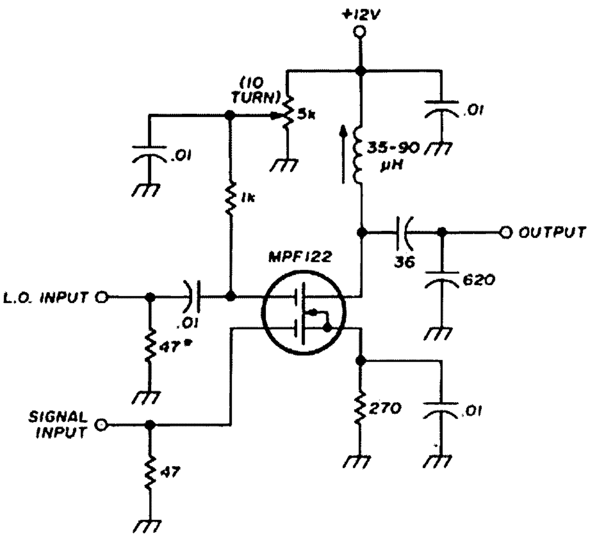


fig. 4. High-frequency mixer circuit using an MPF122 dual-gate mosfet. Resistor marked with an asterisk is removed from the circuit when using a square-wave local oscillator.

I did notice that the input voltage required to saturate the mixer was slightly higher when a square wave was being used as the local oscillator. This was true for both circuits. In general it seems to be possible to replace a double handful of crystals with a frequency synthesizer if some care is taken with the overall receiver design.

references

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2. Signetics Linear Data Book, Volume I, Signetics Corporation, Sunnyvale, California.
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ham radio

understanding spectrum analyzers

Spectrum analyzers are very useful in radio communications — here's how they work and how to use them

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

Spectrum analyzers are instruments which display signal amplitude versus frequency on a cathode-ray tube (CRT). **Fig. 1** shows how a typical spectrum analyzer CRT display might appear. The three vertical responses are produced by three separate input frequencies, and the jagged nature of the base line is caused by the system noise. CW signals appear as vertical lines, and modulated signals will show sidebands.

The Heathkit SB-620 Scanalyzer is a special-purpose spectrum analyzer which is used in conjunction with a communications receiver. It displays all signals whose frequencies are within a few hundred kHz of the frequency to which the receiver is tuned. Thus the operator may visually observe the activity on the band and see where the strong signals and clear frequencies are located without having to tune his receiver.

operation

Fig. 2 is a simplified block diagram of a spectrum analyzer. It is nothing more than a receiver whose frequency is swept across a certain band, and whose output causes a vertical deflection on the CRT for each input signal encountered.

The input network may vary according to system requirements. It may contain a filter to limit the input signals to the desired frequency range, and thus prevent image responses. An amplifier may also be included to improve sensitivity.

The mixer produces an i-f output whose frequency is the difference between the input frequency and the local oscillator frequency. This output is amplified by the i-f amplifier, detected, and fed to the vertical deflection plates of the CRT.

As is the case with an ordinary receiver, the frequency of the spectrum analyzer is controlled by varying the frequency of the local oscillator (LO). Since the spectrum analyzer must automatically and continuously sweep across a band of frequencies, the LO must be voltage controlled so that its frequency may be varied by changing its control voltage.

A sawtooth waveform generator pro-

vides the control voltage for the LO and the sweep signal for the horizontal deflection plates of the CRT. This arrangement causes the electron beam in the CRT to move horizontally across the face of the CRT as the frequency of the LO changes. Thus the horizontal position of a vertical response on the CRT is directly related to frequency. For good accuracy it is important that the sawtooth waveform be linear, and that the relationship of control voltage to LO frequency also be linear.

As an example of how the spectrum analyzer performs, assume the input frequency range is 3 to 8 MHz, and the center frequency of the i-f amplifier is 9 MHz. The LO must sweep from 12 to 17 MHz to produce a difference frequency of 9 MHz between the LO and input. If an input signal exists at 4 MHz, an i-f signal will be produced when the LO frequency sweeps through 13 MHz.

i-f bandwidth

It is desirable that a CW signal appear as a straight vertical line on the CRT. To approach this condition, however, the i-f bandwidth must be very small compared to the frequency range swept by the spectrum analyzer. In the example above, the horizontal scale of the CRT would cover 3 to 8 MHz, a sweep width of 5 MHz. In this case, an i-f bandwidth of 5 kHz would give good resolution because it represents only 0.001 of the total horizontal scale. The CRT response to a CW signal would be quite narrow, being only 0.001 of the total horizontal scale at the 3-dB points.

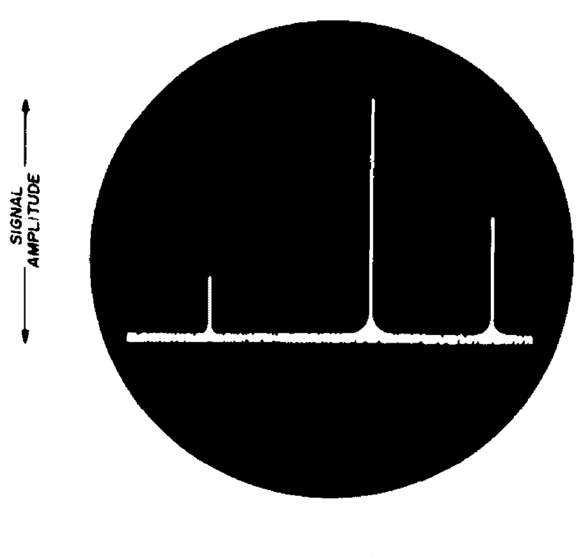


fig. 1. Typical spectrum analyzer display (CW signals).

If, however, the spectrum analyzer is adjusted so that the total frequency range swept is from 3.00 MHz to 3.05 MHz (a sweep width of 50 kHz), then a 5 kHz i-f bandwidth would be only one-tenth of the horizontal scale. This would produce a poor display such as that shown in fig. 3, and the display would show the frequency response of the i-f passband rather than the desired vertical line. This illustrates why the ratio of sweep width to i-f bandwidth should have a value of at least 100, and preferably higher.

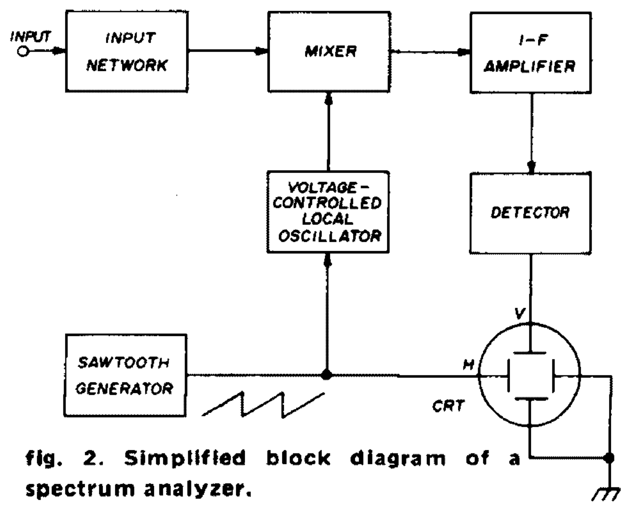


fig. 2. Simplified block diagram of a spectrum analyzer.

sweep speed

As the i-f bandwidth is decreased, it lowers the limit on how fast the frequency range may be swept. This is because the rise time of the i-f amplifier increases as its bandwidth is decreased. If the spectrum analyzer sweeps by a signal too rapidly, the i-f amplifier won't have time to respond to it.

Another problem encountered when the i-f amplifier frequency response has steep skirts is that ringing will occur if the sweep speed is too fast. This will produce a distorted display such as that shown in fig. 4. It can be shown mathematically that, if the i-f passband is determined by

a single tuned circuit, the minimum i-f bandwidth should be

$$\text{i-f } BW_{min} \geqslant 1.18\sqrt{\frac{\text{sweep width}}{\text{sweep time}}} \quad (1)$$

This is seldom if ever the case, because a single tuned circuit would not provide sufficient selectivity.

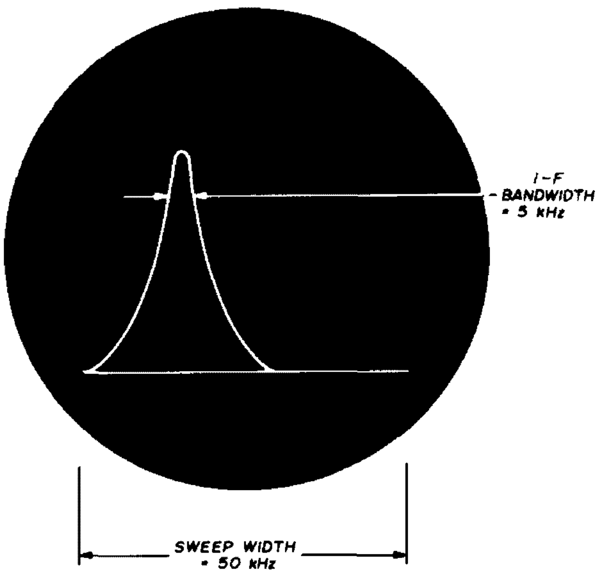


fig. 3. Poor resolution of the spectrum analyzer display occurs when the i-f bandwidth of the instrument is too large a fraction of the sweep width.

It has been determined empirically that if the i-f passband has a nearly ideal rectangular shape, the minimum i-f bandwidth should be

$$\text{i-f } BW_{min} \geqslant 10\sqrt{\frac{\text{sweep width}}{\text{sweep time}}} \quad (2)$$

As an example, assume the horizontal scale of a spectrum analyzer CRT is to be 500 kHz wide, and the time of one sweep is to be 0.01 second (this corresponds to a sawtooth frequency of 100 Hz). The minimum i-f bandwidth is found from **eq. 2** to be

$$\text{i-f BW} = 10\sqrt{\frac{500{,}000}{0.01}} = 70.7 \text{ kHz} \quad (3)$$

Obviously the i-f bandwidth is too wide to give a satisfactory display, because it is **14** percent of the total horizontal scale. To obtain a reasonable display, the sweep speed must be decreased to accommodate a suitable i-f bandwidth. Assuming that a 5-kHz i-f bandwidth will be narrow enough, **eq. 2** may be rearranged and solved for sweep time.

$$\text{sweep time} = \frac{100 \text{ (sweep width)}}{(\text{i-f BW})^2} \quad (4)$$

$$= \frac{100(500{,}000)}{(5000)^2}$$

$$= 2 \text{ seconds}$$

This corresponds to a sawtooth frequency of 0.5 Hz. With such a slow sweep speed, a CRT having long persistence phosphor becomes desirable. It should be remembered that somewhat faster sweep speeds may be used as the i-f passband skirts become less steep.

conclusion

This brief article only barely introduces you to the subject of spectrum analyzers and some of the more important design constraints and criteria, but it is hoped that it will assist those of you who are interested in designing, building or modifying these versatile devices. Spectrum analyzers exist in the form of complete instruments and as attachments to be used with oscilloscopes. In the latter case, the oscilloscope should have a direct-coupled (dc frequency response) horizontal channel to preclude distortion of a low-frequency sawtooth waveform before it reaches the horizontal deflection plates of the CRT.

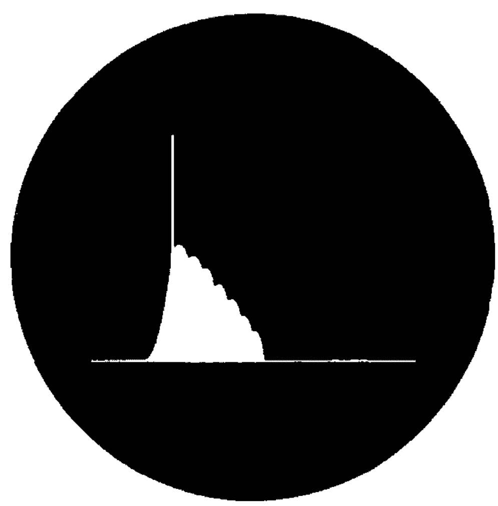


fig. 4. Spectrum analyzer ringing, shown here, is due to excessive sweep speed.

ham radio

adding private-line to the Heathkit HW-202

The Heathkit HW-202 fm transceiver presents special problems if you want to install private-line — here's how to solve them

Arthur Reis, WA8AWJ, 603 N. Court Street, Howell, Michigan 48843

When you live in an area where low-frequency private-line tone is the key to the local fm repeaters, a new transceiver is a challenge. It's not always easy to find the right place to insert the low-audio tone in the existing circuit without creating additional problems. However, amateurs who live in "private-line" areas know that it must be done. And, as more and more repeaters go to private-line, as some have predicted, the challenge will become more widespread.

Recently, Heath introduced its HW-202 two-meter fm transceiver. It's an excellent rig, but it presents some unusual problems to private-line installations. This article shows how to solve these problems.

the circuit

In most fm rigs, the normal practice is to put the private-line circuit (usually a reed vibrator or twin-T circuit) in the modulation chain, *after* the clipping, limiting and deviation adjustment circuits. This is often accomplished by running the output of the private-line board through an appropriate dropping resistor, to the output arm of the deviation control, which is usually only one or two components removed from the modulation diode or transistor. This keeps private-line level independent of any other modulation.

Unfortunately, the Heath HW-202 design doesn't allow for installation of the private-line output at that point. Heath

follows up the output of their modulator (Q202-Q204) with an RC filter chain of high series resistance (R248, R252-R255, C253-C257) (see fig. 1). A 4700-ohm resistor (R256) connects this chain to the modulation diode, D207. Unhappily, any bypassing on the output side of the private-line board (mine is a Motorola CTS) represents a comparative short to

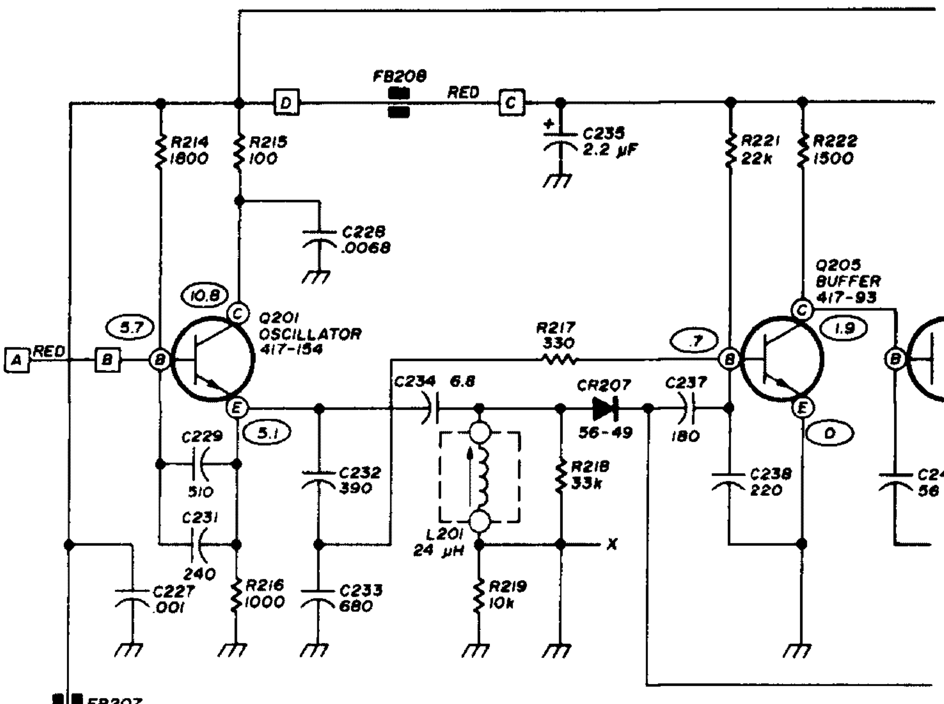


fig. 1. Private-line circuit is connected to the junction of R218, R219, C236 and L201 (point X) in the Heath HW-202.

ground for the modulator output. Deviation drops to zero. It's obvious that another installation point had to be located.

The point finally chosen was more convenient than you might think. Heath inserts the audio to the cathode end of the modulation diode. However, since the diode is not directly grounded, it is possible to modulate either or both sides of the diode at once. Since it was impossible to modulate the cathode in this case, the private-line was inserted on the anode side. Fig. 1 shows the exact point of insertion (junction of L201, R219, R218, and C236). I used a Motorola CTS Board in my rig, which provides 0.5-volt rms output and requires no series resistance of its own for approximately 500-hertz deviation.

installation

Mechanical installation was rather simple. There are no holes to drill if the private-line board is installed in an enclosed space with foam surrounding it. Heath made it especially convenient by putting an extra hole in the transmitter circuit board printed-circuit foil at the junction of L201, R218, R219 and C236. Fig. 2 shows the position of the unused hole on the circuit board.

The output of the private-line board is run through *shielded* wire to that hole. The shield is grounded to the underside of the board at the point marked X *(fig. 2)*. Voltage for the board can be taken from any 13.8-volt point in the unit. I personally chose taking the voltage off the hot side of the pilot lamp on the regulator/hash-filter circuit board.

If hash filtering is needed for the private-line board, as it was in my case, a 220-ohm series resistor, followed by a 500-μF capacitor to ground, will do the job. It may also be advisable to install a switch for the B+ supply to disable the private-line circuit when you're traveling out of town. At least a few repeaters in the Midwest will reject signals containing

private-line tones. Instead of defacing your unit, it may be better to replace the present 2000-ohm squelch control with a push-pull switch type. I set mine to turn off the private-line in the "pulled" position. The squelch function is, of course, not affected.

The private-line unit is wrapped in electrical tape and mounted in foam

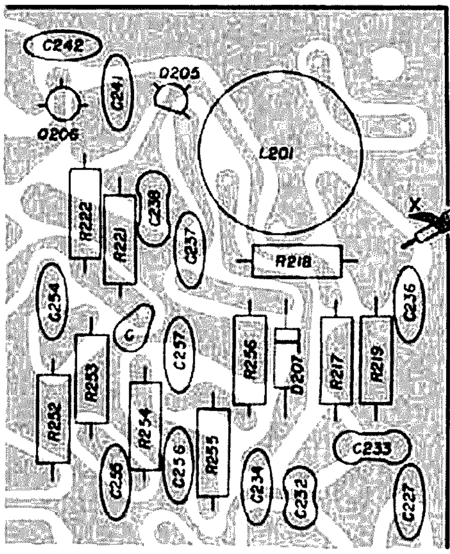


fig. 2. Upper right-hand portion of HW-202 transmitter printed circuit board. Center conductor of shielded cable from the private-line circuit is connected to a hole already in the circuit pad. The shield is connected to the grounded circuit trace.

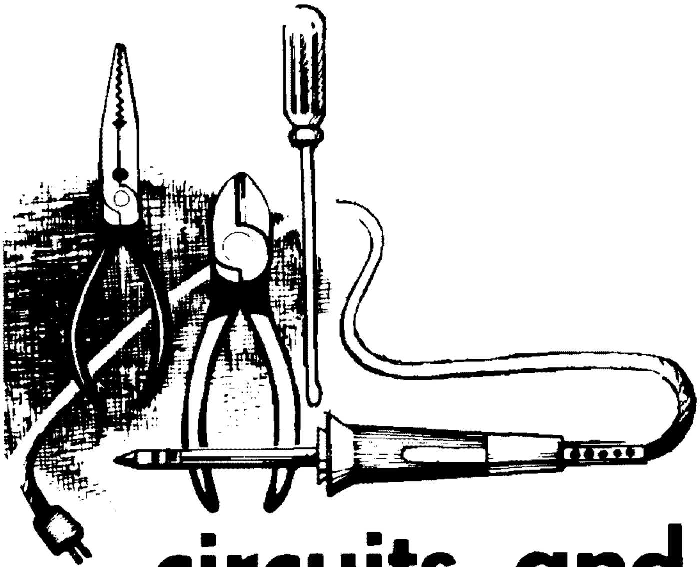
rubber behind the microphone input. Do *not* put the reed near the speaker magnet — reeds and magnets don't mix! The foam rubber is a good shock mount and eliminates the need for any holes in the HW-202.

summary

Performance of my unit is excellent. Private-line level is within one dB of nominal (550-Hz deviation) and there is no interference to the other forms of modulation used in the HW-202 (tone burst and touch-tone, as well as voice).

In conclusion, I would like to especially thank Keith Peterson, W8SDZ, member of the DART Repeater in Detroit, without whose help this project would not have been nearly as successful as it was.

ham radio



circuits and techniques

ed noll, W3FQJ

the dipole beam

Dreaming about beams makes no contacts. Commercial beams can be both expensive and troublesome. Don't forget that the dipole is a beam antenna, albeit bi-directional. It is not unidirectional like a beam but, then again, side rejection might be considerably better than that obtainable from a poor three-band beam. Often QRM off to the side is more the culprit than signal pick-up from the rear. Furthermore, the dipole provides a direct match to 50- or 70-ohm coaxial cable so you don't have to fiddle with any sort of matching arrangement. Matching arrangements restrict bandwidth and introduce loss.

A dipole for 10, 15 or 20 meters is a simple lightweight affair that can be rotated by arm-power or the smallest of TV antenna rotors. Your basic support can be a 10 x 6-inch (25.4 x 15.2-cm) piece of 5/8-inch (16-mm) plastic, fig. 1. Two pairs of U-bolts can be used to hold the two pieces of inexpensive aluminum tubing to the plastic. Another pair of heavier U-bolts holds the antenna to the mast.

Do you need the gain of a large beam? It can be useful for chasing down rare ones. On 10 or 15 meters the band is well open in a given direction or it is closed to sensible communications. A 1 or even 2 to 3 S-unit differential is unimportant in the majority of short DX contacts. The differential is even less important in stateside contacts. Things are tougher on 20 meters but this band is tough for everyone, even those with beams bursting with watts.

Then there is the matter of wind resistance and pocketbook load. The cost per foot of putting your dipole at a specified height is substantially less than that required to put a beam at the same altitude. The dipole is less likely to come down and, if it does, the financial loss or trouble are less than the catastrophe of a disjointed beam.

A split dipole arrangement can be used on more than one band through the use of a line tuner. My preference is for the T-network type (refer to the January, 1973, issue of *ham radio*, page 59). I have used one such 15-meter dipole for almost a year and have worked out pretty much where I wished to on 10 and 15 meters. The radiation pattern is still bidirectional, a standard figure-eight on 15 meters, extended and sharper figure-eight on 10 and a broad figure-eight pattern on 20. If a 20-meter dipole is used, the standard figure-eight pattern is obtained on 20 while a reasonable figure-eight pattern persists on 15. On 10 meters the pattern becomes a lobed affair.

hanging in there

If your dipole is mounted in an accessible location, such as immediately over a chimney or roof top, short sections of insulated wire can be hung on the end of a 15-meter dipole whenever you wish to operate on 20 meters, fig. 2. This simple trick permits 20-meter dipole operation in a tight location and where there is inadequate turning radius (attic mount) for a full-length 20-meter dipole. A better plan is to have a pair of extensions (about 6-feet or 2-meters long) that can be telescoped into the dipole ends.

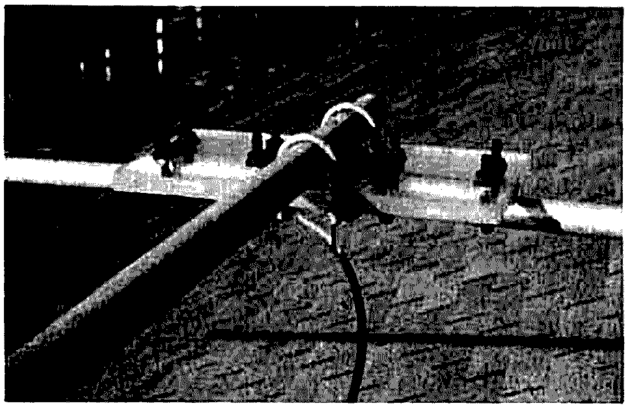


fig. 1. Basic construction of a rotatable dipole element showing the use of U-bolts and a scrap piece of plastic.

15-meter two-element

The same basic construction can be used for a two-element beam, fig. 3. If the reflector is spaced one-quarter wavelength in back of the dipole there is a minimal reduction in antenna impedance. Again, the T-network line tuner is a fine crutch if your transmitter is anemic and requires a perfect match (one of my transmitters is so inclined).

A line tuner permits this same antenna to function very well on 10 meters. The 15-meter reflector is not too long to provide some gain and directivity on 10. You can also load it on 20 meters with a tuner. If readily accessible, telescoping or hanging end pieces can be attached to the 15-meter dipole to resonate it on 20 meters, fig. 4. In this case the 15-meter reflector does some 20-meter *directoring*. Ingenuity is a fun substitute for high costs.

matching emancipation

The matching dispute has two avenues of thought. There is the complete-match phobia and the other. Personally I pursue the other. In most amateur practice antenna-to-line mismatches up to 3-to-1 have little effect on antenna system performance.

Where the mismatch hurts is at the transmitter. This is a matter of transmitter design and is a form of technical enslavement that hampers amateur antenna experimentation. Transmitter mismatches result in loss of output, rapid aging of the output stage, introduction of distortion components (sometimes) and other defects that could be avoided with the design of a wide-impedance output matching system.

In summary, the major ill effects of mismatched loads are a transmitter limitation and not one of antenna performance. The average amateur can overcome this limitation with the use of a tuner. The *manufacturer* can overcome the problem with more versatile design or by including a tuner as part of the transmitter. In modern operation a tuner is not a bad idea at all because of its ability to suppress harmonics and other spurious radiation.

fig. 2. To add 20-meter operation to your 15-meter dipole simply clip an extension wire on each side.

This problem has become very apparent to me because of the inquiries I have received with regard to matching the triangle antenna which has become increasingly popular on 40 and 80 meters

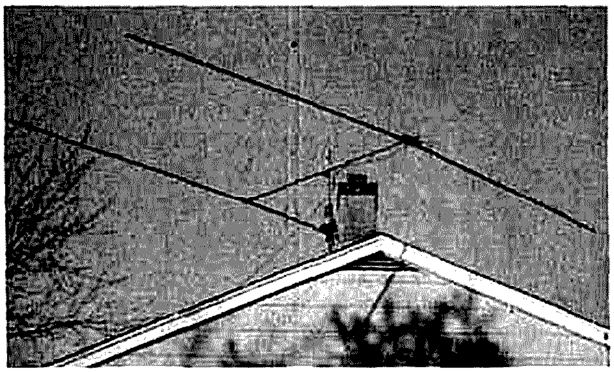


fig. 3. Simple arrangement for adding a reflector to a 15-meter dipole.

and is used on 20 to 160 meters as well. The question usually evolves around obtaining an exact match using a single triangle-driven element. The problem is more prevalent on 20 and 40 meters than on 80 and 160. On the latter two bands the average height of the configuration in terms of wavelength and feedpoint above ground *is* such that the impedance is inherently low once the antenna is resonated.

In the usual forty-meter installation the swr usually falls somewhere between 1.5:1 and 2.5:1. Such a mismatch has no significant effect on the performance of a triangle antenna but is a matter of concern to those worried about swr and possible influence on their transmitter. The swr at the transmitter can be reduced with the use of a 4-to-1 balun and experimentation with the overall length of the transmission line. The ultimate answer is the use of an antenna tuner.

The mismatch problem is a general one with full-wavelength closed antennas such as quad, delta and triangle. Tri-band beams can be an agonizing experience for the amateur striving for that perfect match although antenna performance itself is little affected by what is normally considered a serious mismatch.

In most amateur high-frequency antenna installations the practical length of the transmission line is such that very little loss is generated by the mismatch. Using high-quality low-loss coaxial line or open-wire types it is an insignificant quality.

The full-wave closed antenna performs well at low mounting heights, can be readily positioned and shaped to fit the mounting site and provides good performance despite necessary physical distortions away from straight-line mounting (see the May, 1973, issue of *ham radio*, page 66).

add 160 to your 80-meter inverted dipole

Many backyards that can handle an 80-meter inverted dipole are too confining for the construction of a full-size 160-meter antenna. However, by accepting some distortions away from the straight-on construction you can add 160-meter performance to your 80-meter antenna. The ends of your inverted dipole are low and in most situations it is practical to clip add-on sections that will provide 160-meter resonance. These can

fig. 4. Telescoping section converts a 15-meter dipole to 20 meters. On 20 meters the 15-meter reflector acts as a director.

be clipped on whenever 160-meter operation is desired, producing a very fine performing antenna. They can be run straight away, keeping them 7 feet (2 meters) or more above ground to permit pedestrian traffic. However, resonance can also be obtained by running these

extensions at various angles to meet your property line, fig. 5. I have operated one successfully on 160 meters using the arrangement of fig. 5B.

To calculate the length of the extensions use the quarter-wavelength

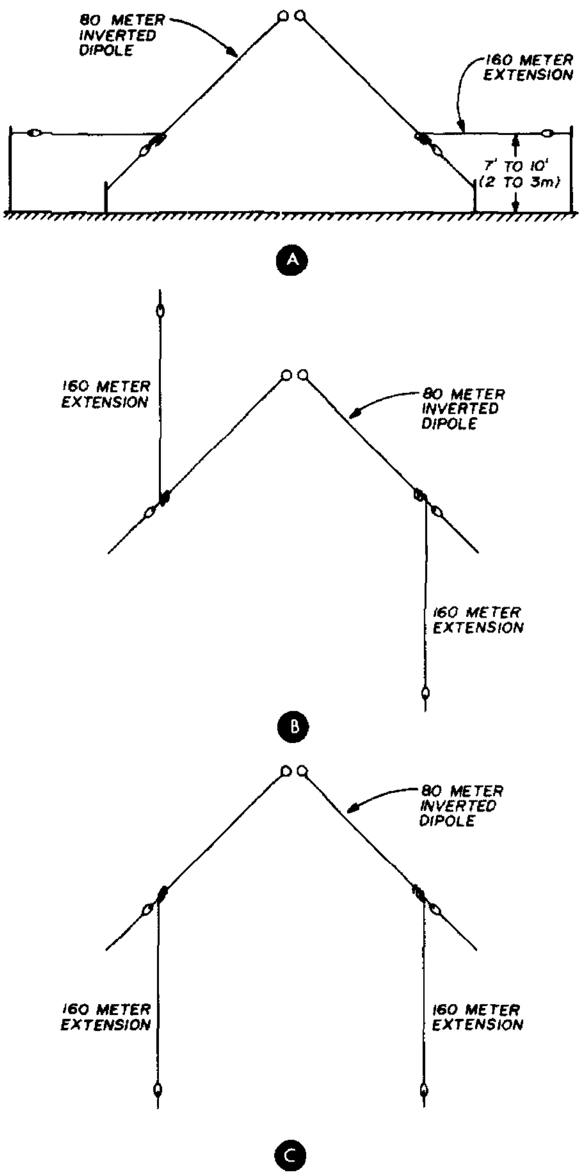


fig. 5. How to add 160-meter extensions to your 80-meter inverted dipole (see text for calculating extension length).

equation, substracting from it the leg length of your 80-meter inverted dipole.

$$L_{ext} = \frac{234}{F_{160}} - L_{80}$$

where L_{ext} is the length of the extension in feet, F_{160} is the desired operating frequency on 160 meters and L_{80} is the length of the 80-meter inverted antenna. This calculation may give you a longer overall length than required, but you can trim back both ends to establish resonance on the desired 160-meter frequency. Usually the extension needs to be cut back further as you fold the legs away from the straight-away plane of your 80-meter inverted dipole.

Using a T-network tuner I was able to load this antenna on every band, 10 through 160 meters. The performance of the configuration of fig. 5C was interesting in that it provided some additional directivity for operation on 10, 15 and 20.

ham-metrics

The English-speaking radio amateur has been talking metrics since the "below 200-meters" days. However, in calculating antenna length amateurs make the unnatural conversion to feet using the appropriate constants instead of going directly to antenna length in meters. The purchase of a meter stick or meter rule provides an easy introduction to the metric system when used in conjunction with antenna calculation. Typically, meter rules come in lengths of 10, 25 or 50 meters.

What is the wavelength of a 4-MHz wave? As you know, wavelength equals the quotient of propagation velocity (300,000,000 meters/second) over the frequency in hertz. Therefore:

$$L_{meters} = \frac{300 \times 10^6}{4 \times 10^6} = 75 \text{ meters}$$

What is a half-wavelength at this frequency? A quarter wavelength?

$$\lambda/2 = \frac{75}{2} = 37.5 \text{ meters}$$

$$\lambda/4 = \frac{75}{4} = 18.75 \text{ meters}$$

What is the length of each quarter-wave segment of a half-wavelength dipole cut for 4 MHz considering an "end effect" of 4%?

$$\lambda/4 = 18.75 \times 0.96 = 18 \text{ meters}$$

basic equations

Three useful equations for making free-space wavelength calculations and one for calculating the length of one segment of a half-wave dipole are as follows:

$$\text{Length (free-space full wave)} = \frac{300}{f_{MHz}} \text{ meters}$$

$$\text{Half wavelength} = \frac{150}{f_{MHz}} \text{ meters}$$

$$\text{Quarter wavelength} = \frac{75}{MHz} \text{ meters}$$

$$\lambda/4 \text{ dipole element} = \frac{72}{f_{MHz}} \text{ meters}$$

For example, to find the length of a quarter-wave segment of a half-wavelength, 80-meter dipole for use on 3.6 MHz, the calculation is as simple as this:

$$\lambda/4 = \frac{72}{3.6} = 20 \text{ meters}$$

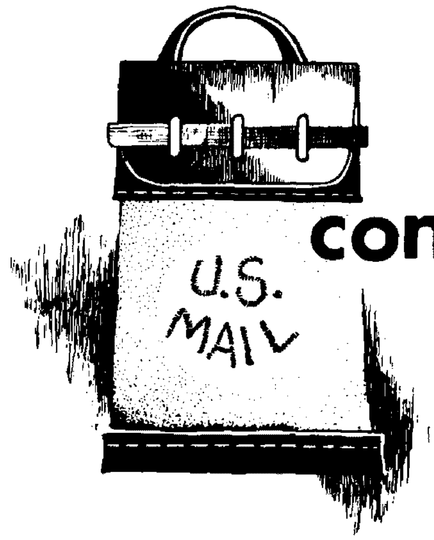
Get out your meter rule and cut your dipole. If, out of curiosity, you may want to know how long your antenna is in feet, you can use a conversion factor to determine this length. Why bother? You have already cut your antenna to resonate at 3.6 MHz. Using the metric system is not so much accepting the new as it is throwing away the old.

Recall the conversion you must make to inches when your answer is in decimal parts of a foot? No such foolishness is required with the metric system. Just remember that there are 10 decimeters in one meter and 10 centimeters in a decimeter (100 centimeters in one meter). All of which can be read directly from meter rules of various types. For example, calculate the length of each segment of a half-wave dipole for operation on 14060 kHz.

$$\lambda/4 = \frac{72}{14.06} = 5.12 \text{ meters}$$

Measure off two lengths of 5 meters, 12 centimeters and you have your 20-meter dipole.

ham radio



comments

solid-state tube replacements

Dear HR:

I was highly interested to see the letter from Mr. Walter Loomis printed in the October, 1973, issue of *ham radio.* The company that I am employed with is in the business of manufacturing semiconductors, and one of our products is very similar to the Fetron that Teledyne produces.

Being active in vhf myself, and having numbers of tube-type communications receivers that use 6AK5 tubes, I gave our devices a try. The following observations may be useful to some of your readers.

1. Our first tube substitute was made in an epoxy block. The performance of this unit was slightly less than could be obtained by going to a ceramic substrate, as Teledyne had done. The main problem with the epoxy package was feedback capacitance, measured at about 0.05 pF.

2. The input to the device is the gate of a jfet. Gate currents of good jfets are in the picoamp range, while *grid* currents in vacuum tubes are in the hundreds of nanoamps, even when operated at several volts of negative bias. This may cause problems with biasing, since the vacuum-tube oriented engineer may have been counting on this electron interception current to provide some operating bias for the tube. The Fetron will not show this bias.

3. Input capacitance of the Teledyne device is specified at 8 pF maximum, which is twice what the tube will have. This will lead to all sorts of impedance-matching troubles, even if the tuned circuit can be resonated, at operating frequencies above approximately 50-MHz.

4. The gate of the input jfet looks like a diode connected from grid to cathode. This diode cannot be forward biased to much more than 0.8 volts. Consequently, applications depending on short pulses of high current (class-C frequency multipliers) cannot be expected to work well, as the tube replacement cannot be enhanced as the original vacuum tube could.

5. Many communications receivers use a small *positive* bias voltage applied to the bypassed end of the grid bias resistor for the rf amplifier. This voltage becomes a few volts negative on strong signals from agc circuitry in the i-f of the receiver. However, when the receiver is idling with no signal input, the positive bias voltage will flow through the resistor and forward bias the gate diode in the input jfet. This bias causes the jfet to show a marked decrease in g_m, usually to about 1/10 of the original value. Thus, a sort of "inverse agc" action results. Additionally, the input impedance decreases markedly under these conditions, further aggravating the low gain problem. The fix here is to short out the agc line.

I have tried these solid-state tube replacements in every receiver that I have using a 6AK5. The results were most encouraging. Our epoxy model tube replacement worked well at 6 meters, showing a 2-dB noise figure improvement over the tube. Gain was slightly higher (about 1 dB) than the tube. Equally good results were obtained in 70-MHz i-f amplifier service. At 147 MHz, results were not

encouraging. The input circuit could not be made to tune to the operating frequency, and overall sensitivity was extremely poor. Switching to our ceramic substrate type of tube replacement did not provide too much improvement at first. However, an investigation of the problem revealed that good results could be obtained.

The noise figure of these devices is quite good, as the input transistor is a somewhat improved 2N4416. The feedback capacitance is extremely low *when the metal can is grounded.* Measurements show it to be approximately 0.01 pF under operating bias. The Y_{os} at the plate terminal is extremely low, about 0.05 micromho. This is much better than the tube. In short, this device can run circles around the tube it replaces, provided the circuit is adjusted to optimize the design of the stage for the replacement device.

After changing around the input and output matching networks of one receiver on 147 MHz, I obtained a sensitivity of 0.2 microvolts for 20-dB quieting (with a 20-kHz i-f bandwidth). Now I will never have to replace my rf stage again and I can expect long and trouble free maintenance of the sensitivity. But I had to work on the receiver to take advantage of the performance of the device.

I have also used the device in low-frequency applications such as limiters at 455 kHz. The tubes replaced here were 6BH6s and 6CB6s. In most of the places I tried them, the solid-state replacements worked as well as the original tubes. In some sockets, instability and oscillation was the result. The short between pins 2 and 7 is probably responsible for this.

An important point about fets that will probably limit the usefulness of solid-state tube replacements in a-m equipment (including ssb) is that the fet is a square law device, while the vacuum tube is only a poor approximation to square law, at best. Why is this? Because an a-m system almost always requires agc, and the agc action of a remote-cutoff pentode is very slow. Useful amplification can be obtained at −10 volts of grid bias with the remote-cutoff pentode, while the square-law jfet will long since have pinched off and become merely a low value capacitor connecting the input and output. On the other hand, since the jfet is such a good square-law device, it will provide excellent intermodulation distortion performance, which is a key requirement in today's equipment.

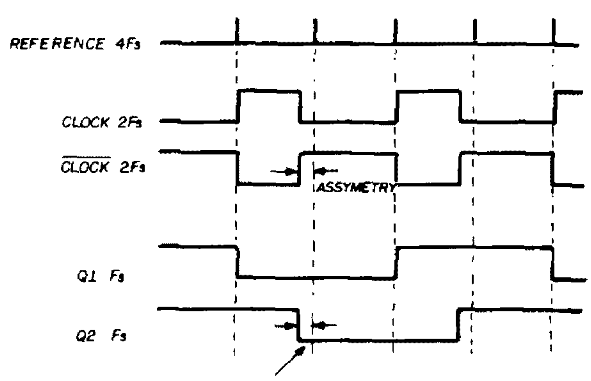
To summarize, I found the solid-state tube replacement to be a great device, but it is also one whose *requirements must be* understood to make it useful.

Bob Hale, WB6APU
Solitron Devices
San Diego, California

phasing receiver

Dear HR:

The Phase II communications receiver described by WAØJYK in the August, 1973, issue is a good example of the practical realization of well-known concepts that can only be achieved after a certain level of technical refinement is reached in the state of the art. Simple concepts often seem the most difficult to implement. I would like to add a few comments that I hope will lead to a better understanding of the digital phase-shifting technique.



The broad-band digital phase shifter used by WAØJYK, while only requiring a two-times frequency division instead of the more common four-times division, depends heavily on symmetry of clock transition timing to maintain phase quadrature. That is, for proper operation, the clock *transitions* must be periodic at 4-times F_{signal} since the basic two-stage shift or switch tail counter *must* have a clock component at four-times F_{signal} to

deliver quadrature phase output at F_{signal}. The paraphase clock driver described in the article is a form of frequency doubling and causes one flip-flop to change state every clock transition. The timing diagram above is offered to demonstrate this point. Clock asymetry is exaggerated for illustration.

Asymetry of the clock transisions at twice the clock frequency may have several causes. Among them are junction saturation effects (minimal in ECL), nonlinear junction capacitance, gain drifts, unequal path propagation delays and, in the circuit used in the article, comparator threshold and hysteresis. All of these effects are proportional to frequency. Differences in the Q1, Q2 flip-flop clock to output propagation times also add to departures from orthogonality but for the 10131 ECL dual D-type flip-flop used, these effects should be smaller than the clock asymetry problems described above. Phase shifts due to differential flip-flop output loading and mixer differences should also be considered.

In summary, if best use is to be made of the broad-band digital quadrature phase generator technique and optimum performance obtained from the high speed devices available, both flip-flops in the chain must see identical clock transitions. This requires, of course, a fundamental clock frequency of four times the desired signal frequency.

Julius M. J. Madey, W6FAW
Fairfax, California

Mr. Madey has pretty well assessed the problems involved. I think in a word the present design might be summed up as economical. The frequency coverage was already limited by the immediate demands of Air Force Mars and no big effort was made to extend the range above 10 MHz. ECL logic is non-saturating and therefore does not radiate as much noise and doesn't cause as much noise in the power supply as does saturating-type logic. The MC1035P is very economical and has proven to be a real workhorse. The comparator threshold and hysteresis that Mr. Madey mentioned no doubt exist in the MC1035 and enter into the poor performance obtained above 10 MHz. Clock symmetry of the MC1035P at 7 MHz looks very good on a Tektronix 545 but even that scope is not sufficient for a fair evaluation.

The MC1035P presents no particular problem in waveform symmetry below 10 MHz providing that either good crystals are used or a vfo is transformer coupled into the crystal socket with at least 1-volt p-p. With transformer coupling the vfo seems to preserve the symmetry of the vfo signal.

The big problem arises when the 10-MHz frequency limit is exceeded or the clock source is a synthesizer with a dissymmetrical output. Unfortunately, most inexpensive synthesizers and, most of all, those with TTL ICs have dissymmetrical outputs. One local amateur, well versed in digital logic, presented a convincing argument for including the digital phase shifter right in the synthesizer.

Perhaps someone will develop this synthesizer with the quadrature output and the Phase II receiver can be pushed to 20, 15 and even, possibly, 10 meters. It seems the digital phase shifter is sort of the thorn on the road to success. There is no way around it but perhaps we can help each other over it.

G. Kent Shubert, WAØJYK

motorola test set

Dear HR:

I want to congratulate you on Dave Maxwell's fine article, "Test Set for Motorola Radios." I have built it and we are now using it to align USAF MARS Motorola units on a MARS repeater frequency in the Las Vegas area.

Two items of interest. The Amphenol microphone jack may be converted to a Motorola microphone jack by sawing a wider guide slot in the Amphenol unit. The Simpson 1212 50-μA meter has an internal resistance of 5300 ohms. Any meter in the 5000-ohm range may be used.

Oscar Heinlein, W7BIF
Boulder City, Nevada

all-channel frequency synthesizer

Dear HR:

The article, "Inexpensive All-Channel Frequency Synthesizer for Two-Meter FM," by Jim Fahnestock, WØOA, which appears in the August, 1973, issue of *ham radio* could cause people some serious problems.

Several years ago I wrote a technical report* on spurious outputs from mixers, and I wondered about such outputs when I read WØOA's article. I used the programs described in that report to analyze the frequency sets he used, and then wrote a rather simple program in BASIC to do a somewhat more thorough analysis. The results of the latter calculation are enclosed for your information (see table 1 below).

table 1. Possibly troublesome spurious mixer products of WØOA's frequency synthesizer using the 4 x 4 matrix. These products are based on an output at 146.28 MHz, using a fundamental frequency of 6095 kHz and 24-times multiplication (F1 = 8125 kHz, F2 = 2030 kHz).

M	N	Frequency	M+N
1	1	6095 kHz	2
1	7	6085 kHz	8
2	5	6100 kHz	7
2	11	6080 kHz	13
3	9	6105 kHz	12
3	15	6075 kHz	18
4	13	6110 kHz	17

I used an output frequency of 146.28 MHz as a basis since that is our local repeater input frequency. The fundamental frequency for a multiplication of 24 is 6095 kHz. Note that for M + N = 7 there is a spurious output only 5 kHz away. There is another spurious output 10 kHz away from the desired signal for M + N = 8. These are both apt to result in strong spurious outputs from the fm transmitter since there is no way of removing them by filtering.

In the article Mr. Fahnestock suggested that "mixing problems arise" when using a 5 x 5 or a 6 x 6 matrix of frequencies. I did all the necessary calculations for a 6 x 6 matrix and analyzed it with the BASIC program. Interestingly enough, the mixing problems are down by two orders of M + N, and the closest spurious frequency goes out to 7.5 kHz away. This is an extremely doubtful practical solution, but it would be many dB better than the 4 x 4 matrix.

It is possible that these :chemes would work if digital mixers were used, but I am by no means certain since I have no experience with them. It seems unlikely that the author considered them with his emphasis on various filters.

Bill Wageman, K5MAT
Los Alamos, New Mexico

*William E. Wageman, "Analysis of the Spurious Frequency Response of Mixers," Los Alamos Scientific Laboratory document LA-4296-MS, December, 1969.

vhf fm in england

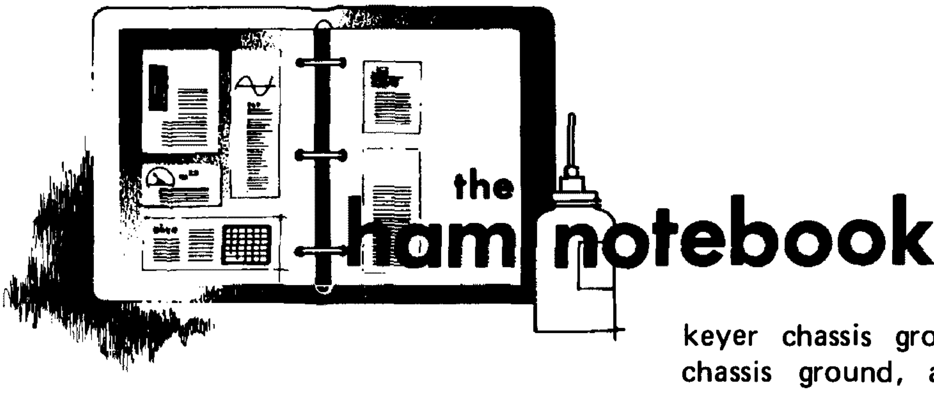
Dear HR:

Here in England vhf fm has really gotten going. We have several fm groups similar to the local one (Southern FM Group). One fm repeater is in operation, licensed as GB3PI, and several more are under construction and waiting license approval.

The Southern Group two-meter repeater with which I am involved will use a modified Storno 613 solid-state base station. The control and ID equipment will be built by club members. Output power will be 25 watts, fed into an Antec antenna (7-dB gain). Input frequency will be 145.125 MHz; output will be at 145.725 MHz. Callsign will be GB3SN.

I have found the repeater articles in *ham radio* to be very useful in designing the control equipment for GB3SN and I would like to thank the American authors. I would be interested to hear from Stateside repeater groups with a view to discussing mutual problems.

Rodney V. Smallwood, G8DGR
11 Wilmot Walk, Wash Common
Newbury, Berks, England



npn transistor switching for electronic keyers

Many amateurs own solid-state keyers which use a pnp transistor switch to key a transmitter's grid-block keying circuit. Such keyers can easily be adapted for use with solid-state transmitters which require a switching positive voltage above ground to negative ground. There are several ways to do this. First, a relay can be inserted between the keyer and the transmitter. Although this method is generally used, the noise of the relay is often found objectionable. Then, too, a relay takes quite a bit of power; this is an important consideration when a keyer is battery-operated.

Second, the pnp transistor in the keyer can be used to do the switching by reversing the leads to the transmitter key jack, but this necessitates separation of keyer chassis ground from transmitter chassis ground, a practice *not to* be recommended. Third, an npn transistor switch can be inserted between the keyer and the transmitter. The keyer would be plugged into the npn switch built into a small Minibox and the latter plugged into the key jack of the transmitter. The keyer, the npn switch and the transmitter would all have a common chassis ground but different electrical grounds. This method is highly recommended. The little device is inexpensive to construct and can be put together in very little time. The schematic and parts list *are* given in **fig. 1.**

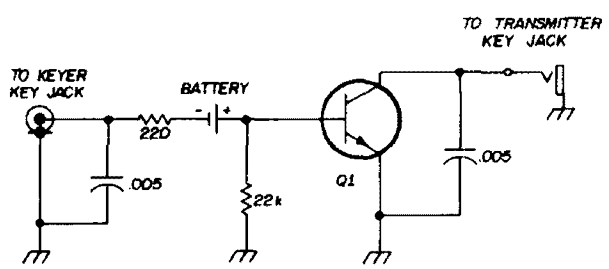


fig. 1. Circuit for keying low positive voltages to negative ground using keyer with pnp switching transistor. Battery is a single penlight call; capacitors are disc ceramics; resistors are ½ watt; Q1 is a Motorola 2N3904 or similar npn transistor.

Those amateurs who have keyers with npn switches but would like to use solid-state switching instead of a relay to control grid-block keying circuits in their transmitters can also use this conversion method. Only two changes in the device described here are necessary to adapt it for that purpose. The polarity of the battery must be reversed and the npn switching transistor in the adaptor must be replaced by a Fairchild 2N4888 or similar pnp transistor.

C. Edward Galbreath, W3QBO

cleaning files

If you do much work with aluminum you will find that your files and cutting tools rapidly plug up with aluminum filings. These filings are usually quite difficult to remove, especially from the finer toothed files. Don't throw these plugged files out, you can make them look like new again with this simple

method. Just soak the files in a solution of warm water and common household lye. The lye will dissolve the aluminum particles and the hydrogen bubbles released by this reaction will help to dislodge any metal particles that may also be plugging the file. When working with lye it is wise to wear rubber gloves and avoid spilling or dripping the solution on any surface that you don't want harmed.

Pete Walton

build 5/8-wavelength 144-MHz antennas from CB mobile whips

The popularity of 3-dB gain mobile antennas for two-meter fm communications has very little to do with their steep cost. The following is an easy method for converting inexpensive CB whips to 5/8-wave antennas that are identical to commercial gain antennas. In many cases 27-MHz CB whips may be obtained at half the cost or free, depending on your sources of supply.

Before you run over to your local CB supplier, or patronize the want ads, be sure that you can identify the proper type of whip for this conversion. These whips appear to be identical to commercial vhf whips (148-158 MHz), the difference being in the hidden resonator within the epoxy sleeve antenna base. CB whips should dip to around 27 MHz on a grid-dip meter.

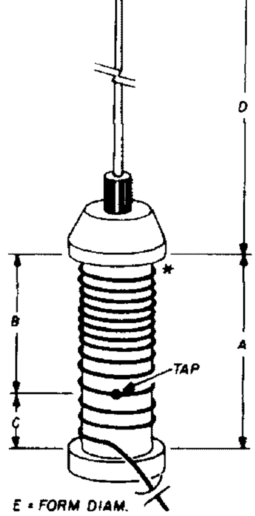


fig. 2. Modifying a CB mobile whip to 5/8-wavelength whip for two-meter fm. No. 12 wire is used for the winding. Connection at top of winding is on other side (point marked with asterisk).

Some epoxy sleeves may be pried loose from the antenna with the aid of a sharp screwdriver. Others may have to be unscrewed by gripping the chrome fittings at both ends with two pairs of pliers. Removing the core should be easily accomplished by friendly persuasion from a hammer and soft wood rod. Getting past the ring of sealer may require some additional effort and patience (five minutes).

Winding the new coil may require five minutes or so of your time. The diameter of the coil forms may vary slightly among manufacturers, but present no problem. Just make sure that your tap remains at three turns above the ground end. The only compensation which may be necessary is extending or shortening the radiating element by a few inches. Two

table 1. Typical dimensions of CB mobile whips (see fig. 2).

	A	B	C	D	E
whip 1	2"	6½ turns (1½")	3 turns (½")	44"	5/8"
whip 2	1½"	7½ turns (1")	3 turns (½")	44"	1/2"

specific examples of typical whips are shown in table 1 to eliminate any guesswork.

Most commercial whips are universal as far as roof mounts, hardware and threading are concerned. Manufacturers make available individual components which are contained within the antennas. By simply screwing together some of this hardware, your modified equipment can be made to accommodate your existing equipment.

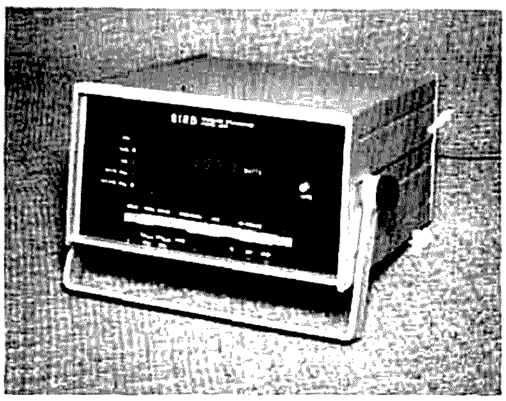
Also, you don't have to limit yourself to available commercial whips. The coil winding data is applicable to plexiglass rods, and readily available coaxial connectors for the homebrewer. You'll find, as I did, that none of the specifications are critical.

There is no noticeable difference between my commercial whip and the modified antenna. Both exhibit significant advantage over quarter wave whips. The swr is 1.05:1.

Robert Harris, WB4WSU

new products

digital in-line rf wattmeter



The new Bird 4371 *Thruline*® Directional High-Power Wattmeter is the first digital insertion instrument for measuring forward or reflected CW power in coaxial transmission lines. It accurately measures power flow under any load condition from 25 to 520 MHz and from 1–1000 watts in six ranges. Insertion vswr in 50-ohm systems is a low 1.1 and accuracy is ±5% of full scale. Model 4371 is also the first high-power directional wattmeter which the user can calibrate in the field to known rf power standards, eliminating weeks of transit for periodic certifications.

The new multi-range digital *Thruline*® Wattmeter measures CW, a-m, fm and ssb signals. No plug-in elements are needed since all variable measurement parameters—frequency range, forward/reflected power and full scale values—are push-button selectable on the front panel. The readout unit and the line-section may be separated by as much as 5 feet for operational convenience. The six power ranges feature 25% over-ranging and high directivity. Bird Quick-Change QC Connectors, which mate with male or female N, BNC, UHF, and others add to the unequalled flexibility of the new RF Wattmeter.

Available from Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139. For more information, use *check-off* on page 94.

RTTY handbook

The new *Teleprinter Handbook* published by the RSGB and available from HR Books is one of the most complete books ever published on the subject. Under preparation by G3LLZ and G3NTT for more than three years, this welcome handbook fills a long-standing need of RTTY operators. The *Teleprinter Handbook* covers all aspects of RTTY communications, including complete mechanical, operating and lubrication data on machines made by Teletype Corporation, Creed and Siemens as well as the surplus TT4 and TG7-A. Information is also included on teleprinters, perforators, reperforators, signal generators and distortion measuring sets.

The chapter on terminal units includes complete construction information for several popular RTTY demodulators including the W6FFC's Mainline ST-5 and ST-6 and TT/L-2. Several commercial units are also described including the popular AN/URA-8B. This chapter also has some good practical information on diversity reception, regenerative repeaters, phase-locked loops and frequency-shift keying. The auxiliary equipment chapter covers such diverse subjects as polar relays, mercury-wetted relays, and electronic and mechanical speed controls. The section on polar relays provides complete operation, maintenance and applications information for a large variety

More Details? CHECK–OFF Page 94

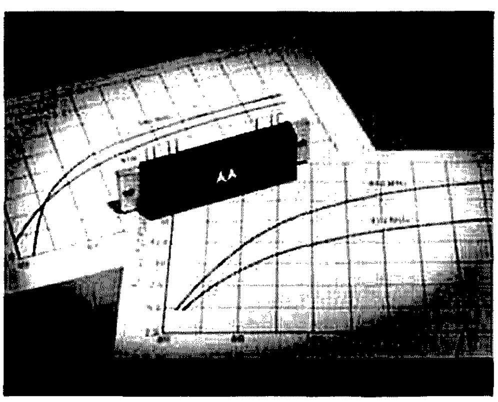
of currently-available polar relays including the Western Electric 215A and 255A.

The chapter on keying discusses FSK and AFSK techniques and includes circuits for several keying units, both commercial and amateur built. Various filter circuits are discussed in detail, with design and construciton information for both passive and active types. The test equipment section provides information on various RTTY measurements, and the control systems chapter tells you how to set up your own RTTY station. The *Teleprinter Handbook* also covers RTTY operating procedures and contests.

One of the most useful sections in the *Teleprinter Handbook* is the glossary of commercial equipment, which provides a short rundown on practically every unit of available RTTY equipment. The appendix provides much practical data including complete design information for simple LC filters based on 88-mH toroids, gear speed identification and teleprinter compatibility.

This book is probably the best teleprinter handbook ever published, and combines a lot of hard-to-find RTTY information all in one place. Highly recommended for RTTY enthusiasts. Hard cover, 13 chapters plus appendix and index, 324 pages, $14.95 from HR Books, Greenville, New Hampshire 03048.

432-MHz hybrid rf amplifiers



Designed for complete rf amplifier or driver applications in uhf land mobile communications systems, two new rf

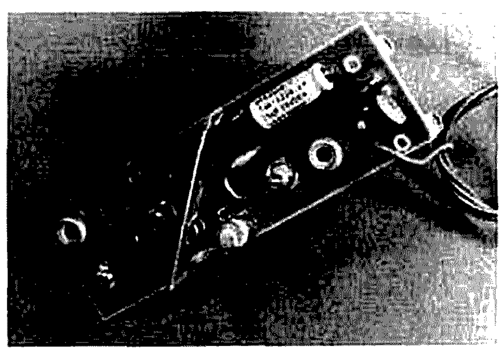
amplifier modules introduced by Motorola offer more than 18 dB power gain at 432 MHz. Designated the MHW709, (7.5-watt) and the MHW710 (13.0-watt) uhf power modules, these are complete amplifier units capable of covering the 400 to 470 MHz frequency range.

Both units operate from a 12.5-volt dc supply which is common in most mobile applications. The MHW 709 delivers 7.5-watts output with a driving power of approximately 100-mW for a power gain of 18.8 dB. A full 13.0-watts can be produced by the MHW710 with only 150-mW of driving power; this represents a 19.4-dB power gain. The frequency range is covered in two bands with the units designated MHW709-1 and MHW710-1 for 400 to 440 MHz, and MHW709-2 and MHW710-2 to cover the 440 to 470 MHz range.

Harmonic suppression is at least -40 dB down across the frequency range with all spurious outputs more than 70-dB below the desired signal. Input impedance is 50 ohms for both modules, and operation with a 20:1 load mismatch produces no damage to the unit.

For more information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 94.

450-MHz preamp



Topeka FM Engineering has introduced a new 450-MHz preamp that features a dual-gate mosfet in the front end. The mosfet provides superior cross-modulation performance and reduced spurious responses. The new preamp, the model 450M, is built on low-loss epoxy

More Details? CHECK—OFF Page 94

circuit board, and has a voltage gain of approximately 15 dB with a typical noise figure of 4.5 dB. The circuit board is silver plated for maximum efficiency, and is shielded on both sides for maximum isolation of the input and output circuits.

The 450M preamplifier is designed to operate from 10 to 15 volts dc, but may be used with higher supply voltages when the accessory HF450PK adapter is used. Similar preamplifiers are available for 406 to 470 MHz. The preamplifier is priced at $29.95, complete with instructions and mounting hardware. The HF450PK power supply adapter is $1.25. For more information, write to Topeka FM Engineering, 1313 East 18th Terrace, Topeka, Kansas 66607, or use *check-off* on page 94.

hard-to-find electronic components

Radio Shack has added more than 2,000 hard-to-find electronics items to its parts and test equipment line with a special "Qwick-Fill" electronic parts catalog. The 52-page Radio Shack Electronic Parts Catalog is available on request from any Radio Shack store. It includes a variety of special-purpose tubes, transistors, readouts, PC and IC equipment, relays, resistors, capacitors, potentiometers, transformers, sophisticated test instruments, connectors, power supplies and other items previously unavailable from Radio Shack or other electronics stores in most localities.

Through a special arrangement with Allied Electronics, Radio Shack's sister company, orders may be placed at any of more than 1800 participating Radio Shack stores. Allied Electronics is known as one of the nation's leading electronic parts suppliers for business, industry and scientific users. There is no minimum order requirement and only a $1.00 shipping and handling charge. Orders are delivered to the Radio Shack store for pickup by the customer. For more information use *check-off* on page 94.

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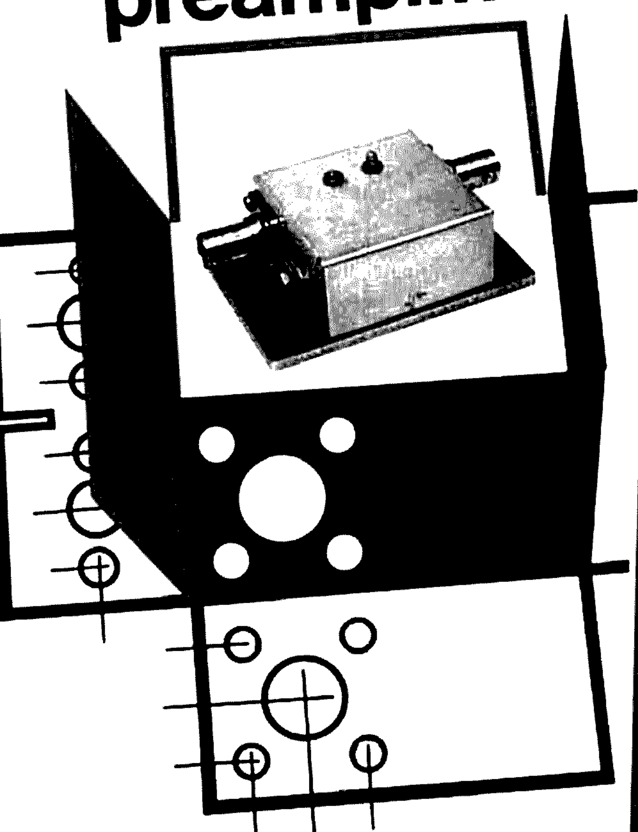
ham radio

magazine

JULY 1974

narrow-band solid-state 2304-MHz preamplifiers



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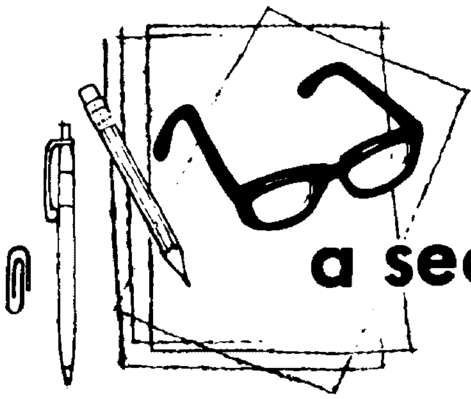


a second look

by jim fisk

Widespread rumors that restructuring of the amateur service was being studied by the FCC were confirmed on May 10th at a joint meeting between FCC officials and ARRL Directors and officers.* In as much as the specific details of the proposed restructuring are still being studied by the FCC staff, and must be presented to the Commissioners before a formal proposal can be issued, the discussion was only in very general terms.

The goal of the proposed plan is apparently to both broaden and upgrade the amateur service, encouraging potential amateurs to join the ranks while at the same time encouraging individual amateurs to improve their operating and technical skills. This means that there will probably be more classes of amateur licenses in the future, including a no-code version to stimulate newcomers, as well as extensive modifications to the amateur licenses themselves.

The proposed restructuring was only one of several topics discussed at the lengthy, all-day meeting. Also included on the agenda was a discussion of the World Administrative Radio Conference scheduled for 1979, and the formation of a National Amateur Radio Advisory Committee. Since other radio services already have such advisory committees, and have for some time, the idea is new only as it applies to amateur radio. Pending approval by the Commission, the first meeting could take place as early as September.

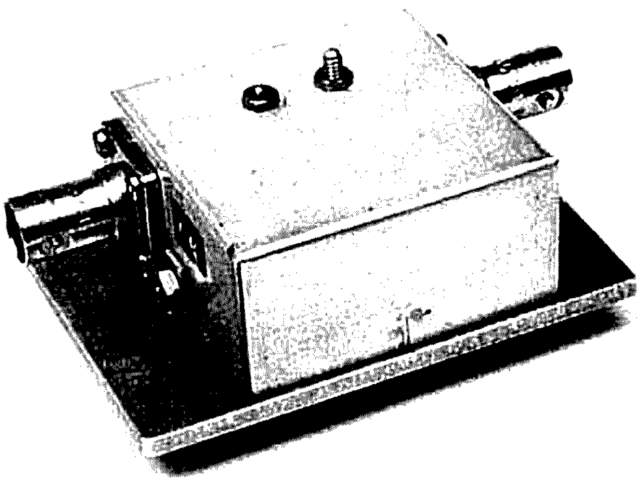
The proposed Committee, which would give the amateur community a much-needed opportunity to work more closely with the Commission, would be chaired by an FCC official and would consist of 12 to 15 appointed members. The members, selected to represent various segments of amateur radio, would meet periodically to advise or make recommendations to the FCC.

With the specter of a World Administrative Radio Conference looming on the horizon only five years from now, this Advisory Committee could be particularly important. A Spectrum Planning Subcommittee Working Group on the Amateur Services (designated SPS WG/A, for short) has been meeting in Washington since early this year, and they have proposed adding 200 kHz to the 40-meter band, 150 kHz on 20 and 50 kHz on 15. They have also proposed *new* amateur bands at 10.1 - 10.6 MHz, 18.1 - 18.6 MHz and 24.0 - 24.5 MHz. With communications satellites assuming more and more of the burden of long-distance commercial and government traffic, these enlarged high-frequency amateur allocations are a distinct possibility.

However, at the 1979 conference the United States will have only one vote, as will each of the smaller, emerging nations, so it will take a lot of work to muster the necessary support to make even part of this become reality. Also, any pressure that satellites relieve on our high-frequency allocations will probably be forcefully reasserted on our vhf and uhf bands. French amateurs have already lost the exclusive use of 144-146 MHz (which must now be shared with the military), and other nations are known to be eyeing the vhf amateur bands. To meet these challenges, and be successful, will require a great deal of preparation — preparation that must begin *now*.

Jim Fisk, W1DTY
editor-in-chief

*A complete report on this meeting is available from *ham radio* and will be sent upon receipt of a self-addressed, stamped envelope.



narrow-band solid-state 2304-MHz preamplifiers

Complete construction details for bipolar 2304-MHz preamplifier circuits featuring 6- to 9-dB gain and 2.5- to 4.5-dB noise figures

Norman J. Foot, WA9HUV, Elmhurst, Illinois 60126

Since the publication of several simple solid-state 2304-MHz converters in the amateur radio magazines,[1,2,3] interest in the 2300-MHz amateur band has grown by leaps and bounds. Recent solid-state devices include bipolar transistors which work effectively at S-band (1550-4000 MHz). It is now possible for the amateur uhf enthusiast to build a 2304-MHz preamplifier using any one of a number of available devices. The virtues of adding a preamplifier ahead of the mixer are well known and will not be repeated here.

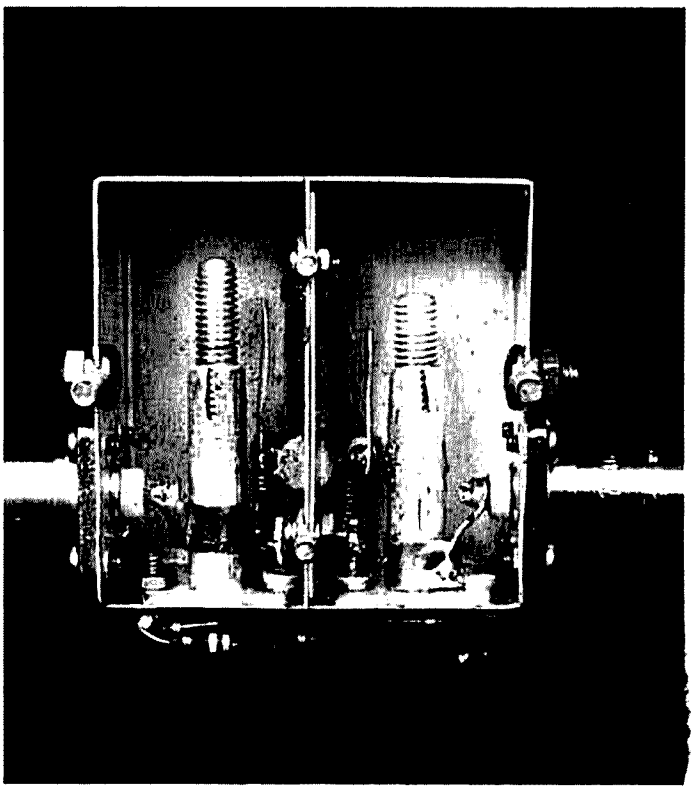
Most 2300-MHz preamplifiers which have been described in recent years feature the well known broadband strip-line circuit.[5] It is also possible to use these transistors in narrow-band circuits which include input and output tuned circuits of the kind normally associated with the lower-frequency bands. One can present a good case for choosing the narrow-band approach where strong signals from nearby TV or fm stations cause receiver desensitization.

The narrow-band 2304-MHz preamplifier described here has excellent frequency selectivity because it uses cavity resonators at both the input and output. It provides 6- to 9-dB of gain and exhibits a noise figure between 2.5 and 4.5 dB, depending on the particular transistor. Four of these 2304-MHz preamplifiers have been built, each using a different type of transistor, including the Fairchild MT-2500 and MT-4500,* and the Hewlett-Packard HP-35821E and HP-35862E. The Nippon Electric V912 also works well in the narrow-band circuit. The performance of these devices is compared in table 1. When compared to a parametric amplifier, these devices give a good accounting of themselves.

preamp design

The electrical design of each amplifier was based on the published scattering parameters of the bipolar transistor used in the circuit.[6,7,8] Fixed values of complementary input and output reactances were used to achieve the desired impedance matching. It is not necessary for you to become involved with these details, so long as the circuit presented here is faithfully copied. This is particularly important in view of the fragile nature of these microwave transistors—they are not very forgiving when circuit changes are attempted.

Top view of the 2304-MHz preamp using a Fairchild MT-2500 transistor. For more details see fig. 1.



construction

The photograph shows the completed amplifier. It is built in a silver-plated brass box 7/8-inch (22-mm) high, 1-3/4-inch (44-mm) wide and 1-3/4-inch (44-mm) deep. Details of the construction are shown in the mechanical drawing, figs. 3 and 4. The bottom of the box is sanded flat after fabrication so that the cover will make good electrical contact all around. Before the transistor is mounted in the enclosure its emitter lead (or leads) are soldered to a brass plate; then the plate is screwed to the partition so that the base and collector leads project into their respective cavities. These leads in turn are soldered to small Teflon standoff insulators to prevent the leads from being damaged when working down inside the close quarters of the box with the soldering iron.

The Fairchild MT-2500 used in the preamplifier shown in the photographs is a stripline-type device with three leads (see fig. 1). The TI-line package, also used in an otherwise identical amplifier, has a pair of emitter leads diagonally opposite one another which allow somewhat better mechanical attachment.

Type BNC connectors (UG-290A/U) are used for the rf input and output connectors. The input is tightly coupled to achieve a low noise figure while the output coupling is adjusted for maximum gain.

The center cavity conductors are made

*Although both the MT-2500 and MT-4500 have been discontinued by Fairchild Semiconductor, often these devices can still be obtained. A somewhat similar, although improved, device in a small 100-mil-square package that should work in the same narrow-band 2304-MHz circuit is the currently available Fairchild FTM-4005. Editor

of ¼-inch (6-mm) OD brass tubing, 1-1/8-inch (29-mm) long, with 10-32 threads on the inside. Tuning is accomplished by running 1-inch (25-mm) long 10-32 brass screws in and out of the open end of the center conductors with a screwdriver. Each screw is slotted on one end for this purpose. The open ends of the center-conductor tubing are slotted with a fine hacksaw blade and then pinched together slightly like a collet so the tuning screws fit tightly.

The base lead of the MT-2500 (or MT-4500) transistor is coupled to the input cavity by means of a brass strip mounted parallel to the center conductor, one end of which is soldered to the base. This circuit resonates at approximately 2304 MHz. For the HP-35821E, a very small 5-pF dipped-mica capacitor is used instead, soldered between the base lead and the midpoint of the center conductor (see **fig. 2**). In both designs tuning of the input cavity is rather broad.

The dc return from the transistor base is through a 1/8-watt, 1000-ohm resistor to a 470-pF Allen Bradley FA-5C 471W feedthrough capacitor. The pigtail of the 1/8-watt resistor is wound into an rf choke of two or three turns before it is soldered to the base. This also provides strain relief to the transistor.

table 1. Typical performance, at 2000 MHz, of microwave transistors suitable for use in the narrow-band 2304-MHz preamplifier. Note that recommended collector current for minimum noise figure does not coincide with maximum gain, or vice versa.

transistor type	collector current	noise figure	gain	price
MT-2500	2 mA	4.3 dB	6.5 dB	NA
MT-2500	3 mA	4.7 dB	7.2 dB	NA
MT-2500	6 mA	5.5 dB	7.7 dB	NA
MT-4500	3 mA	5.4 dB	—	NA
MT-4500	5 mA	5.5 dB	7.5 dB	NA
MT-4500	9 mA	6.0 dB	8.3 dB	NA
FMT-4215	5 mA	4.2 dB	10.0 dB	$17.50
FMT-4575	5 mA	2.5 dB	12.0 dB	$44.00
FMT-4005	3 mA	2.0 dB	12.0 dB	$70.00
FMT-4005	5 mA	2.2 dB	13.0 dB	$70.00
H-P 35821E	3 mA	4.2 dB	6.5 dB	$20.00
H-P 35862E	5 mA	3.3 dB	11.0 dB	$55.00
NEC V912	3 mA	3.5 dB	9.0 dB	$25.00
NEC V912	10 mA	4.0 dB	11.0 dB	$25.00

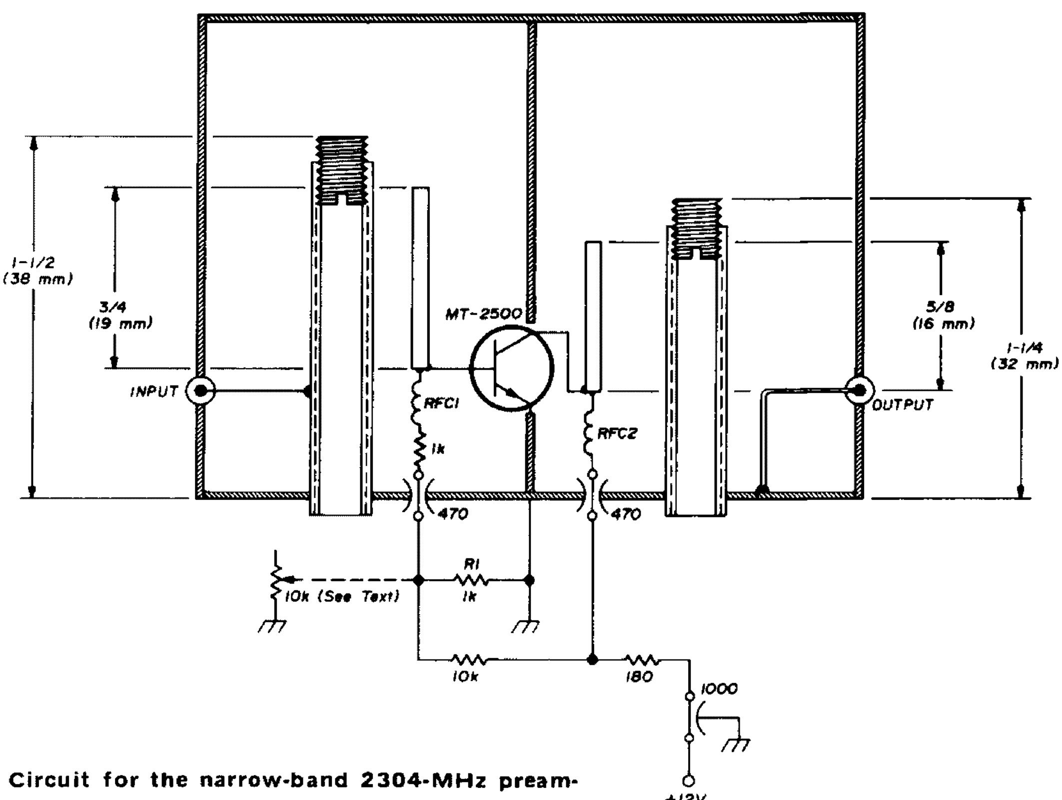
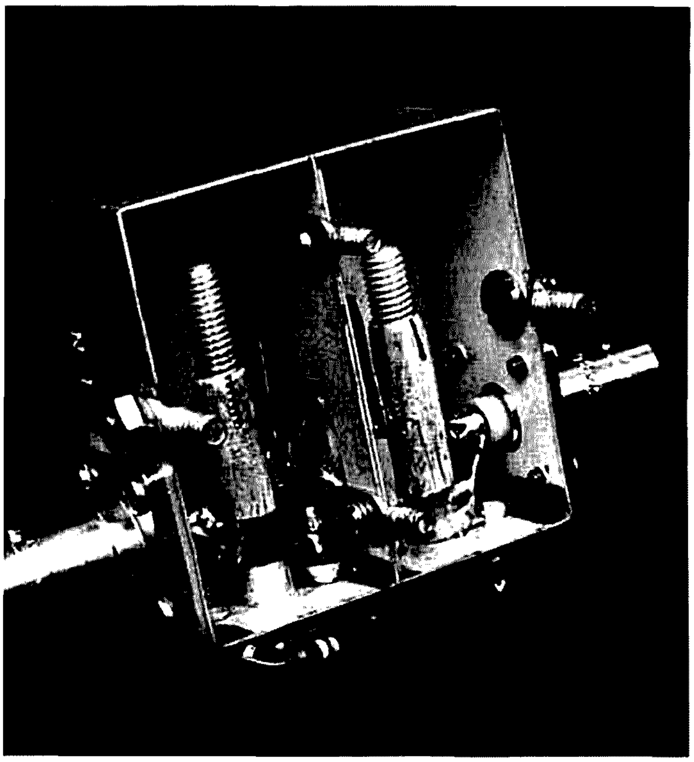


fig. 1. Circuit for the narrow-band 2304-MHz preamplifier using a Fairchild MT-2500 transistor. Similar circuit can be used with the Fairchild MT-4500 or FMT-4005. The 470-pF feedthrough capacitors are Allen Bradley type FA-5C 471W. RFC1 is 3 turns, RFC2 is 5 turns, both no. 26 enamelled wire, airwound using a no. 52 drill as a mandrel (0.0635" or 1.6-mm diameter). Coupling strips on base and collector of transistor are 0.010" (0.25-mm) brass shim stock.



The 2304-MHz preamps are built into small 1-3/4" square chassis made from 0.020" thick brass. Full-size sheet-metal layout is shown in fig. 3.

The collector lead of the Fairchild MT-2500 is coupled to the output cavity by means of a 0.010-inch (0.25-mm) brass strip which is parallel to the center conductor. One end of the strip is soldered to the collector and resonates at a frequency lower than 2304 MHz. The dc return is through an rf choke and another Allen Bradley 470-pF feed-through capacitor. The HP-35821E pre-amplifier works better with a tiny loop coupled to the center conductor as shown in fig. 2.

The 2304-MHz amplifier is fitted with four 6-32 spade screws mounted on the bottom of the enclosure for attaching a bottom cover. As an alternative the pre-amplifier can be mounted on the top of a flat surface such as an aluminum chassis with the use of the spade screws.

tuneup

Before connecting the preamplifier to the 12-volt supply, tack-solder a 10k potentiometer across the 1000-ohm resis-tor, R1. Set the potentiometer resistance to zero (shorted out). Monitor the supply current with a 0-10 mA meter in series with the 12-volt supply. Initially, the meter should register about a half milli-ampere. Now, slowly increase the poten-tiometer resistance (current reading

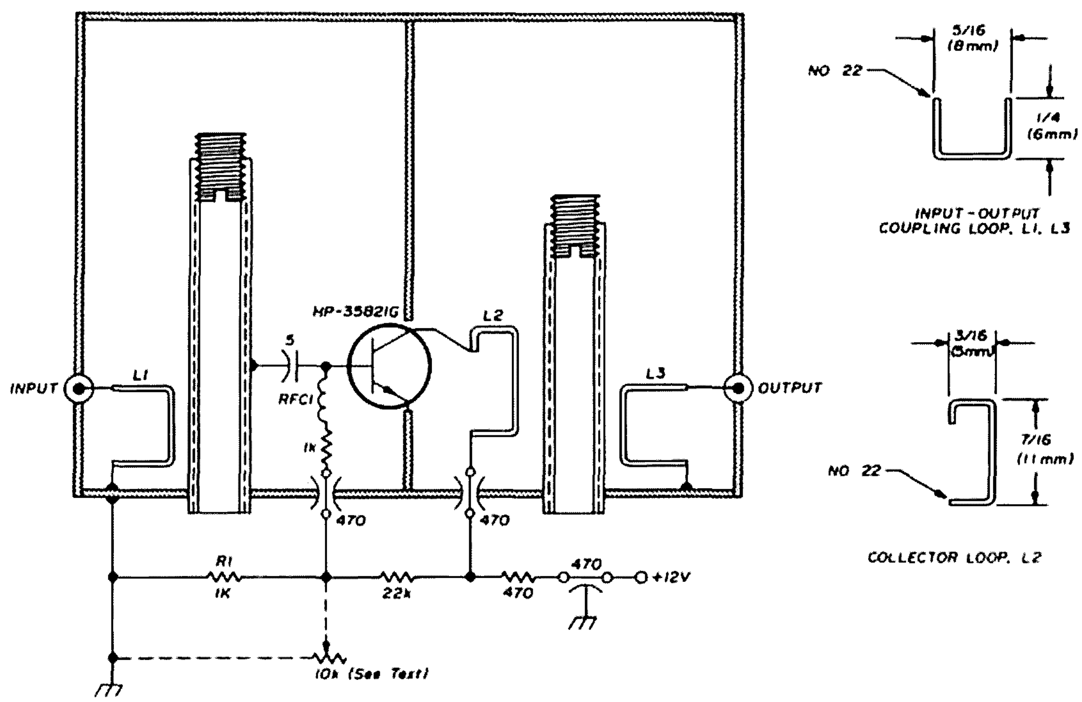


fig. 2. Narrow-band 2304-MHz preamplifier design for the HP-35821E and HP-35862E using coupling loops. The 470-pF feedthrough capacitors are Allen Bradley type FA-5C 471W. RFC1 is 2 turns of resistor lead, airwound using a no. 52 drill as a mandrel (0.0635" or 1.6-mm diameter). All cavity dimensions are as in fig. 1.

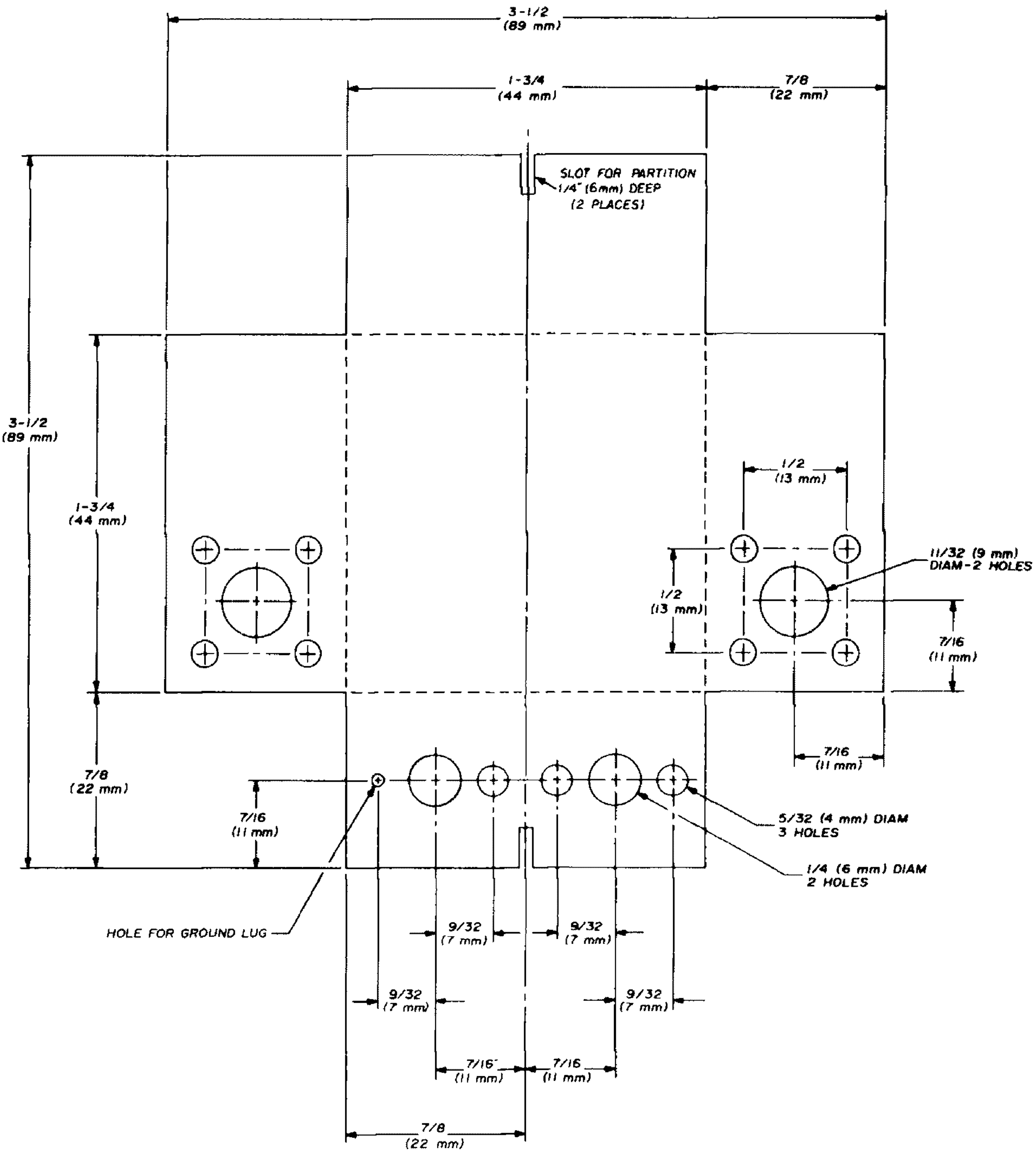


fig. 3. Layout for the enclosure for the 2304-MHz preamplifier. Material is 0.020-inch (0.5-mm) brass or copper, bent along dotted lines. Use a small torch for small solder fillets along all seams. Holes for the spade lugs are not shown (see text).

should increase). If the current exceeds 4 mA before the 10k potentiometer is full-on, back off the current to 4 mA. Disconnect the 12-volt supply and the potentiometer (in that order) and measure the resistance of the potentiometer. Solder a fixed resistor of approximately the same value in shunt with the 1000-ohm resistor.

On the other hand, if the current reading does not reach 4 mA with the potentiometer full on, disconnect the supply and remove the potentiometer. Also remove the 1000-ohm resistor and substitute the 10k potentiometer in its place. Starting with the potentiometer *shorted out*, reconnect the 12-volt supply and slowly increase potentiometer resistance until the meter indicates 4.0 mA. Disconnect the supply and the potentiometer, and permanently solder in a fixed resistor with a value as close as possible to

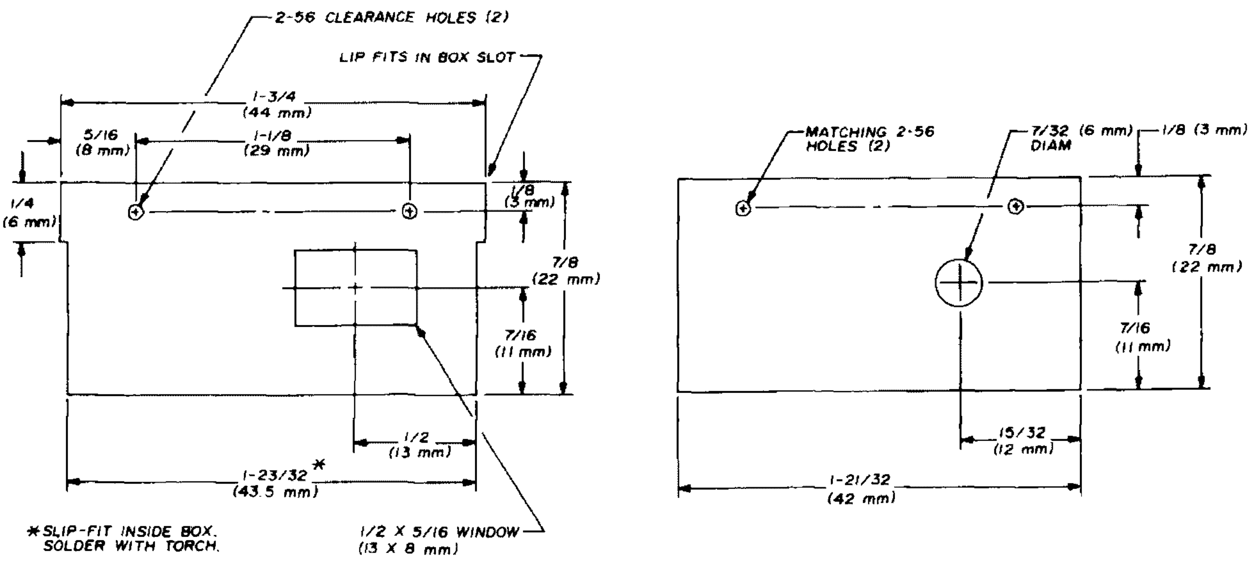
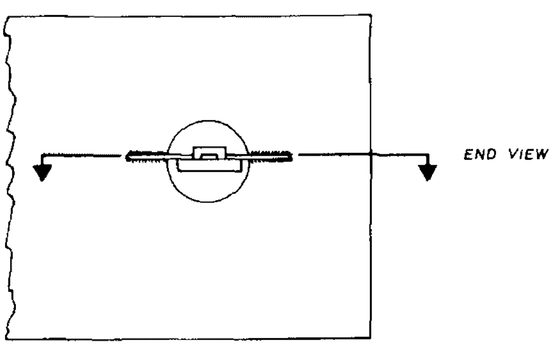


fig. 4. Layout of the center partition and transistor mounting plate. The two pieces are sandwiched together with 2-56 machine screws as shown in the photographs. Material is 0.020" (0.5-mm) brass or copper.

the measured value. Now the preamplifier is ready to go.

At this point there is no substitute for a signal source. With the preamplifier connected to the mixer, adjust both tuning screws for maximum received signal. Adjust the spacing between the collector strip from the center conductor **(fig. 1)** or adjust the collector coupling loop for maximum gain **(fig. 2)**. Check the signal-to-noise ratio with and without the preamplifier—you should be pleasantly surprised.



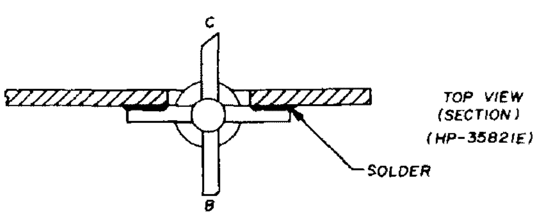


fig. 5. Mounting the transistor on the transistor-mounting plate. After the transistor is installed the mounting plate is attached to the center partition (see fig. 4).

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ham radio

improving the R390A product detector

Eugene A. Hubbell, W7DI, 6633 East Palo Verde Lane, Scottsdale, Arizona 85251

Most improvements in ham gear construction or design are built on a foundation of work done by others. The modification of the military surplus R390A receiver described here owes a lot to articles[1,2] by Captain Paul H. Lee, W3JHR. Some comments by Harry Hyder, W7IV, were also helpful.

Captain Lee's conversion of the R390A to a product detector was tried, and worked satisfactorily except for two details. One of these was a regenerative effect that occurred at the frequency of the bfo, resulting in a peak in the audio response. The other was the loss of the noise limiter for the ssb and CW modes.

Disconnecting the shielded wiring suggested in that conversion and using short direct leads under the chassis to transfer the audio from the a-m detector to the product detector removed the regenerative effect. Restoration of the noise limiter action was not so easy.

When using the noise limiter with the a-m detector a negative bias voltage appears across the combination of R526 and R527 in series and thus, across R120, the limiter control. This voltage results, of course, from signal rectification through the a-m detector, V506B. Captain Lee's circuit omits the limiter entirely in the ssb/CW position as signals will not pass this stage without some negative bias on the cathodes of V507. A check of the similar noise limiter circuit in the Collins 75A4 receiver shows such a voltage switched into the limiter control circuit from the receiver bias supply.

Trying to operate both a-m and product detectors simultaneously without switching outputs did not work out. The product detector bfo is switched on and off from the front panel, but the a-m detector is not so easily disabled. Examination of the a-m detector circuit showed an i-f filter in the transformer T503 lead to R526 and R527. Part of this filter was a 12-mH rf choke. If one end of this choke could be switched from the a-m detector to the product detector it would provide the desired audio transfer, and by introducing a bias voltage in parallel with the product detector signal, the noise limiter problem would be solved.

circuit modification

To control a circuit at a distance, I think of relays. In the R390A, 225 volts is switched on and off from the front panel by the bfo *on-off* switch, and this switched voltage appears at a tie point near the socket for the bfo. Checking a junk-box relay with a 10k-ohm coil, it was found that a 51k resistor in series with the coil would provide reliable operation at sufficiently low current drain to avoid overloading the 225-volt supply.

A negative dc supply for limiter bias in the *ssb/CW* position was provided easily by rectifying the 25.2 volts ac found on the current-regulator ballast tube adjacent to the bfo socket. The last problem had been solved, and here's the step-by-step procedure by which the complete conver-

sion was done. Tech Manual TM 11-856A is almost a positive must!

1. Remove power plug P112 from the i-f chassis along with the three rf plugs P114, P213 and P218. Disconnect the *selectivity* and *bfo* tuning shafts, loosen the mounting screws and remove the i-f chassis from the receiver.

2. Loosen the set screws on the flexible-bellows shaft coupler to the bfo transformer, remove the front shaft and bearing and remove the shaft coupler. Socket XV505 is now more accessible.

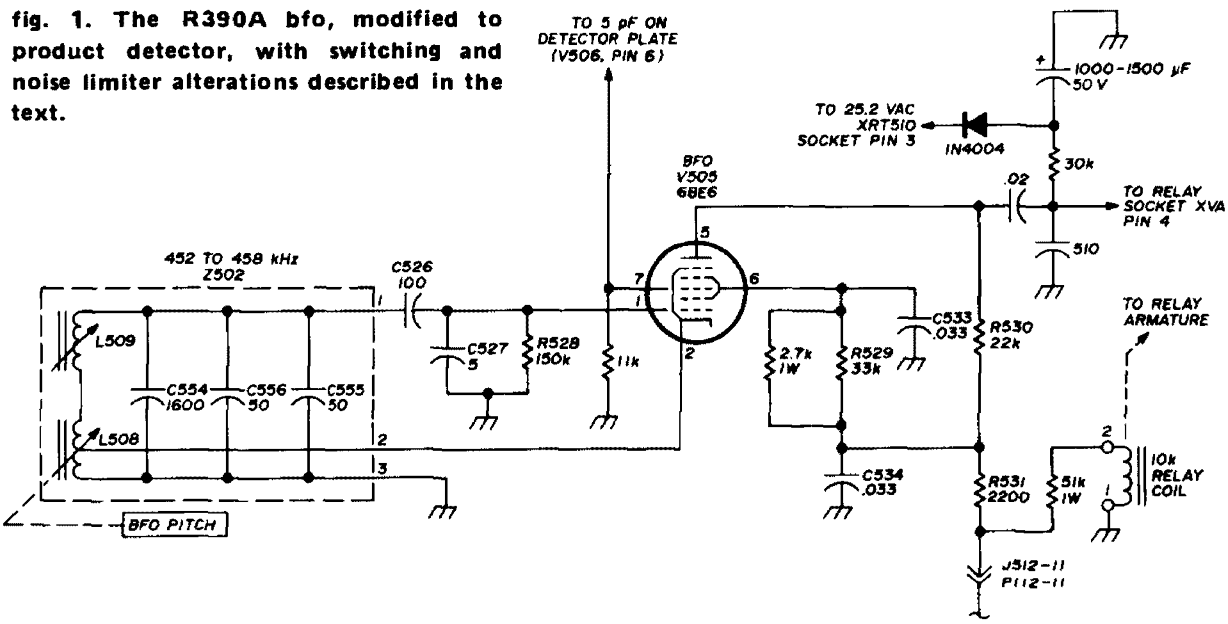
3. Between sockets XV502 and XV506 are three tie points in a triangle. The two

This requires a 5/8-inch (16-mm) hole for the socket and two number-33 holes for mounting bolts. On the chassis below XV503 there should be a dual ground lug holding leads from capacitors C528, 0.1 μF, and C529, 0.033 μF. Remove the nut holding this ground lug and mount the tie point saved from step 3 above.

5. Clear socket terminal 2 of XV505 of all leads. If other leads go to pin 2 for grounding, re-route them to other ground points.

6. Change the cathode tap from the bfo can from pin 7 to pin 2 of XV505. This is lead number 2 on the bfo can.

fig. 1. The R390A bfo, modified to product detector, with switching and noise limiter alterations described in the text.



nearest socket XV502 will not be disturbed. Unsolder resistor R518, 100 ohms, and the connecting wire from the third tie point and remove the tie point from the chassis. Save it. Connect the wire lead back to R518, providing insulating tubing over the joint, and leaving as much slack as possible. Train the resistor close to socket XV503 and the wire between the two remaining tie points and socket XV502. A clear space about 3/4 x 1 inch (19 x 25 mm) should be left between sockets XV502, XV503, XV506 and the bfo transformer can.

4. In the cleared space drill suitable holes and mount a 7-pin miniature tube socket, which we will call XVA, for the relay.

7. Connect a 11K, 1/2-watt resistor from pin 7 to ground.

8. Remove C535, 12 pF, and discard.

9. Connect 2.7k, 1-watt resistor in parallel with R529, the 33k, 1/2-watt screen-dropping resistor.

10. Connect a 5-pF mica capacitor from pin 7 on XV505 to pin 6 on XV506. This couples the i-f signal into the product detector.

11. Solder one end of a 510-pF mica capacitor to a ground lug near XV505 and one end of a physically small 0.02μF capacitor to pin terminal 5 of XV505. Join the two remaining capacitor terminals and attach a wire lead. The other

end of this wire lead connects to pin 4 of the new 7 pin socket, XVA. In doing these last operations near socket XV505, be sure to leave room for the bellows shaft coupling so nothing will be shorted out when the coupling is reinstalled.

12. Ground pin 1 on the new socket XVA. Connect a 51k, 1-watt resistor from pin 7 on this socket to the tie point nearest pins 3 and 4 on socket XV505. This tie point already has a lead and one end of R531, 2200 ohms, 1/2 watt, fastened to it. Insulate the leads on the 51k resistor and dress it next to the chassis to facilitate heat transfer.

regulator tube RT510, and has 25.2 volts ac on pins 2 and 3. Install a 1N4004 diode between pin 2 (cathode end) and pin 6, which also has a wire lead attached at this time. Use an ohmmeter to check that pin six is not connected inside the regulator tube, and is merely used as a tie point here. Connect the other end of the wire lead from pin 6 to the tie point you moved and reinstalled in **step 4** above.

15. Connect a 30k, 1/2-watt resistor from this same tie point to terminal 4 on socket XVA, which already has a wire lead from the output of the product detector. Connect a 1000- to 1500-μF,

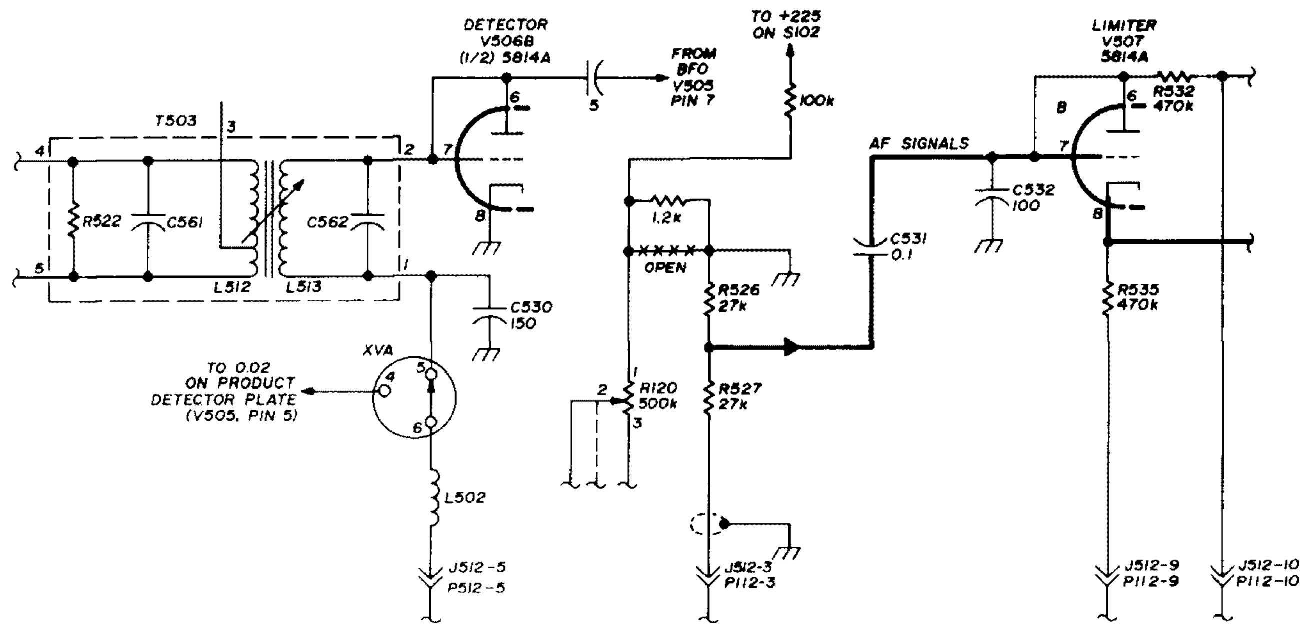


fig. 2. The R390A a-m detector and noise limiter circuits showing changes necessary to use the noise limiter with the product detector.

13. At the front of the chassis, between the *bfo tune* and the *selectivity* switch shafts, there is a molded 12-mH rf choke mounted on a spade bolt. One terminal of this rf choke connects to terminal 1 of transformer T503, which also has a capacitor C530, 150 pF, to ground. Disconnect the rf choke from the transformer, and run a wire lead from the rf choke to pin 6 of socket XVA. From pin 5 of XVA, run a lead to terminal 1 of T503.

14. Socket XRT510 is located in the corner of the chassis below the bfo tuning shaft. This is the socket for current

50-volt electrolytic capacitor from chassis ground (positive terminal) to the tie point just used for the wire lead from the diode. I found a lug under the spade bolt holding the 12-mH rf choke worked out fine for the positive (grounded) terminal.

16. Replace V505, a 6BA6, with a 6BE6. The new socket XVA is for a 7-pin plug-in relay with a 10k-ohm coil. The present Potter and Brumfield number for this unit is PW5LS, I believe, although SM5LS and XSM-1135-2 seem to be the same. About 35 volts of the 225-volt dc supply will appear across this relay coil when the bfo is turned on, switching the

audio input to the noise limiter from the a-m detector to the product detector. The output of the product detector is across the 30k resistor connected between pin 4 of socket XVA and the negative end of the electrolytic capacitor. About 20 volts dc will appear across this resistor. Audio output level should be just about equal for either detector system.

17. To provide a positive cutoff of audio feeding through the noise limiter tube, V507, when the limiter control, R120, is advanced, it is necessary to supply a small positive voltage to the normally grounded end of R120. To do this a small voltage divider must be installed, and it will help if the front panel is partially removed, or at least pulled forward a couple of inches on the tuning and bandswitching shafts. See instructions for panel removal in the Tech Manual. This is to provide access to the back of the *function* switch, S102, and to R120, the limiter control.

18. Remove the ground lead from the grounded end of R120 and the switch on the back of R120. Replace the lead with a 1.2k, 1/2-watt resistor. From the same terminal on R120 connect one end of a 100k, 1/2-watt resistor. The other end of this resistor must connect, either by its own lead or an extension wire, to the terminal on S102, the *function* switch, which turns on 225 volts dc when this switch is in *agc, mgc* or *cal* positions. This switch terminal is just below the ac line switch, a microswitch type with heavy terminals and two white and orange wires considerably larger in diameter than anything else nearby. A check with a voltmeter should confirm that you have found the right terminal.

check out

To check to see that things are going to work, you can reinstall the bellows coupler to the bfo can and the panel shaft bearing and shaft so the bfo can be tuned. Connect the rf input and output couplers, P114, P213 and P218, and plug in the power plug, P112. By placing a box or other support under the i-f chassis, it will be possible to turn it approximately 180 degrees from its normal operating position. This is best done by turning the receiver on end, the i-f end down. Now it should be possible to reach terminals inside the i-f chassis for voltage checks while the receiver is working.

Operating the receiver in this position is a bit awkward, especially turning the selectivity switch and tuning the bfo, but the selectivity positions may be counted from the stop at either end of rotation. With the receiver operating, rf and audio gains turned well up, but with no antenna, a considerable hiss should be heard. Set the selectivity switch to 1 kHz and adjust the bfo tuning for the lowest pitched hiss. This should be equivalent to setting the bfo to *zero* on the front panel.

A check of the voltages on the limiter pot should be made with the vom; one end should register about 16 to 20 volts negative, and the other end about two volts positive. With an antenna connected and any normal noise level, it should be possible to observe the limiting action as the limiter control is advanced in a clockwise direction. If everything checks out ok, shut off the power and put the receiver back together.

summary

I believe this modification is very worthwhile. The product detector action is good, the noise limiter is very good on CW, and the changes have a neat appearance. The diagrams show the changes in the schematic, and the step-by-step conversion is not difficult to make. One warning comment: Make sure your limiter pot does not have an open in it. I had one that was bad, and it really caused me a headache for a while.

The R390A makes a very good second receiver for the shack, is invaluable in the shop, and is really well built. If only it weren't so heavy!

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ham radio

miniature 7-MHz transceiver

Howard F. Batie, W7BBX, 2912 Johnson Road, Falls Church, Virginia

Project shrink — a Quality Recipe for a Pocket Portable

How small can a complete transceiver be made and yet retain enough features to permit consistently reliable operation? Dick Tracy notwithstanding, it seems to come down to the size of the front panel required to mount the controls for the functions desired, and not the space required for the electronic circuitry itself.

To verify this point, "Project Shrink" was undertaken to construct a complete, yet consistently useful, transceiver in as small a case as possible using commonly available parts. By useful, it is meant that the receiver should have a sensitivity on the order of a microvolt and the transmitter be in the 2- to 3-watt input class. Additionally, a vfo or vxo is considered mandatory.

By eliminating all frills, careful attention in the PC-board layout phase, and the selection of a multi-element IC, a very credibly performing transceiver can be constructed using only four solid-state devices; however, the characteristics shown in table 1 attest to the fact that this is a very useful portable or emergency transceiver. The functional block diagram is shown in fig. 1.

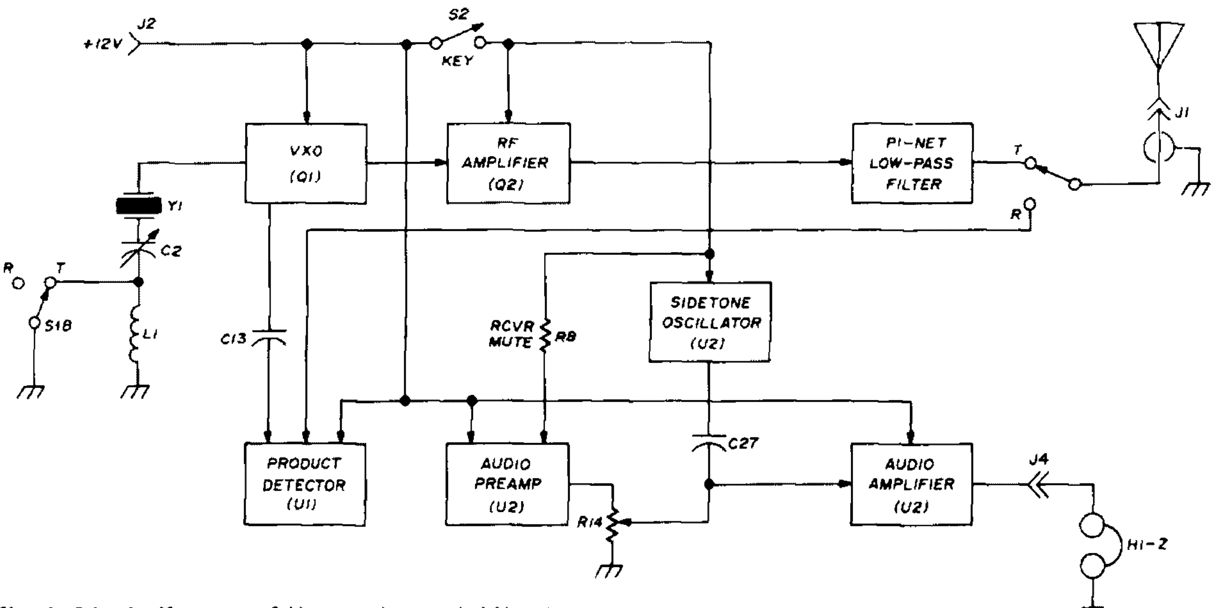


fig. 1. Block diagram of the miniature 7-MHz transceiver.

transmitter

High-beta transistors were used at both Q1 and Q2. The vxo feature was attained by placing a miniature variable capacitor from a transistor radio in series with the crystal, permitting up to about a 4-kHz excursion of the crystal frequency, depending upon the activity of the crystal. A 7-microhenry coil in series with the crystal and variable capacitor was experimentally determined to offset the frequency shift of the oscillator by the correct amount when transmitting. This allows the transmitter during transmit to be zero-beat with the receiver during receive, and yet retain the vxo function.

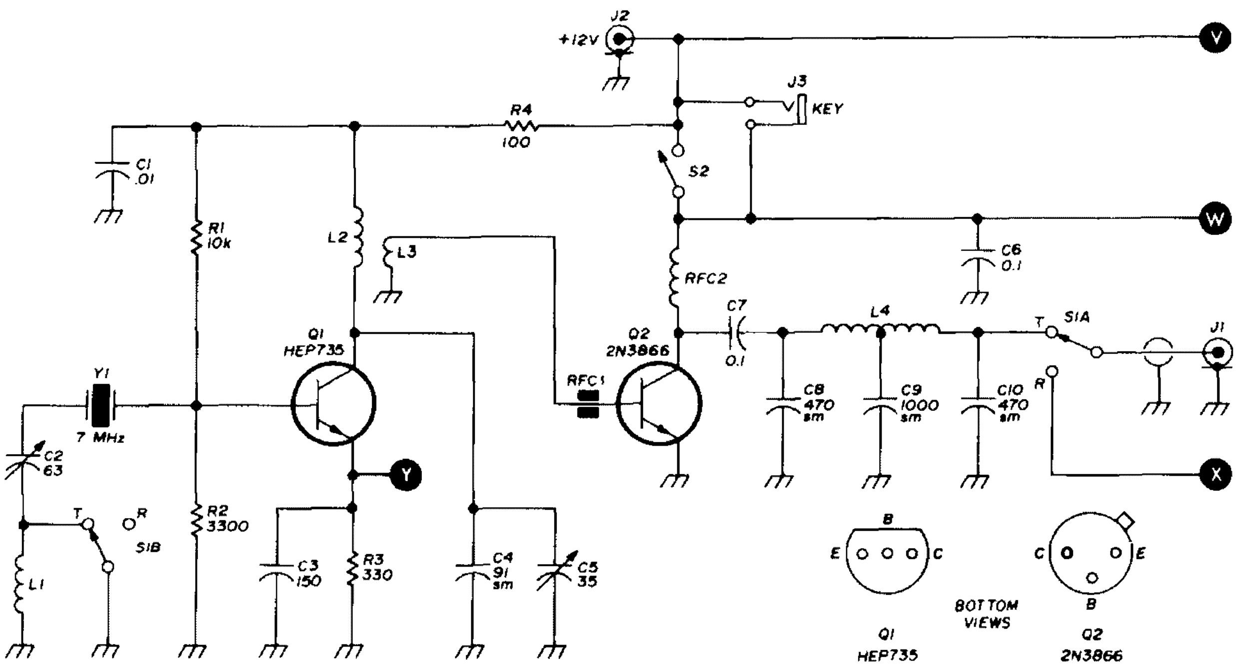
A simple class-C rf amplifier followed by a fixed pi-net low pass filter designed for 52 ohms completes the transmitter. Transistor Q2 collector current is 170 mA at 12.0 volts dc input. A pushbutton key is included in a convenient (and operable) position on the top of the cabinet. Left-handed operators may want to reverse the entire layout in mirror-image fashion. The key should not be depressed during "receive" since no load for Q2 is present; however, Q2 was purposely keyed for 10 seconds with no load and no damage resulted. It should be noted that since the $+V_{CC}$ to Q2 is keyed, any external key must be capable of handling the full Q2 collector current.

table 1. Operating characteristics of the miniature transceiver.

Frequency	7.0—7.3 MHz, vxo controlled
VXO excursion	4 kHz nominal
Transmitter output impedance	52 ohms, unbalanced
Receiver input impedance	52 ohms, unbalanced
Size	1¼ x 4¼ x 2¼ inches, 12 cubic inches (3.2 x 10.8 x 5.7 cm, 197 cc)
Weight	7 ounces (198 grams), including key and five crystals
Number of active devices	2 transistors, 2 ICs

Current drain, 52-ohm resistive load:

	9-volts	12-volts	15-volts	18-volts
Receive	15 mA	21 mA	28 mA	34 mA
Transmit	185 mA	220 mA	240 mA	250 mA
Transmitter input power	1.35 W	2.3 W	3.2 W	3.9 W



C2	BC variable, 63 pF per section (see text).	L3	6 turns no. 28 enamelled wound around L2
C5	3-35 pF trimmer (Arco 403)	L4	20 turns no. 24 enamelled, center-tapped, on Amidon T-50-2 toroid core.
J1,J2	phono jacks (Calectro F2-806)		
J3	3.5-mm phone jack, insulate from chassis.	RFC1	Amidon ferrite bead mounted close to base of Q2
L1	50 turns no. 32 enamelled on Amidon T-37-10 toroid core	RFC2	50-μH rf choke
		S1	miniature dpdt slide switch
L2	38 turns no. 28 enamelled on Amidon T-37-10 toroid core	S2	miniature spst pushbutton switch
		Y1	7 MHz crystal

fig. 2. Schematic diagram of the transmitter used in the miniature transmitter. Transistor Q2 must be provided with a heatsink. In some cases a 75-pF silver-mica capacitor may be required from the base of Q1 to the emitter to sustain oscillation.

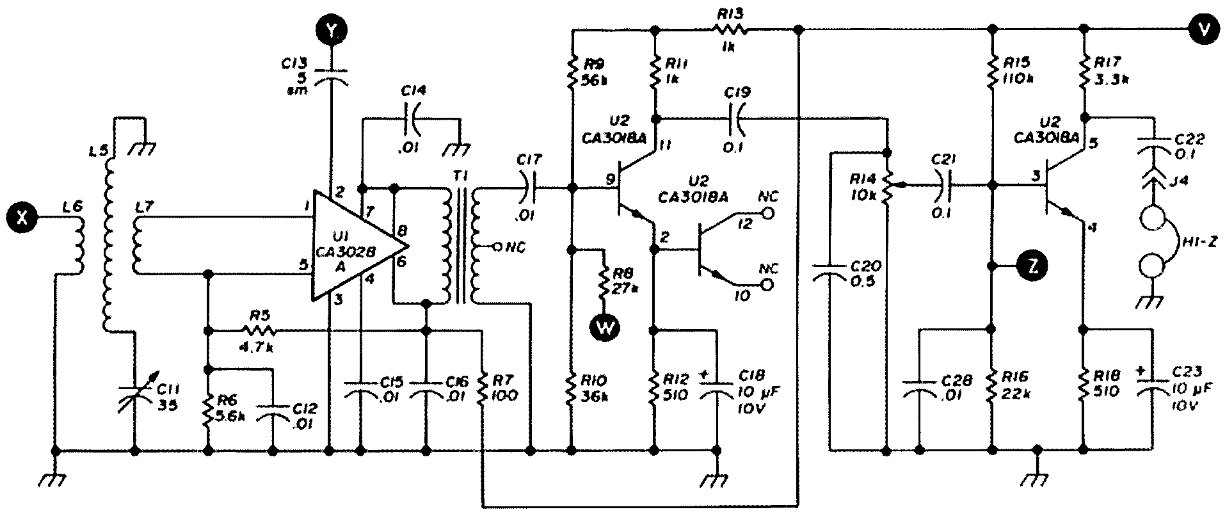
receiver

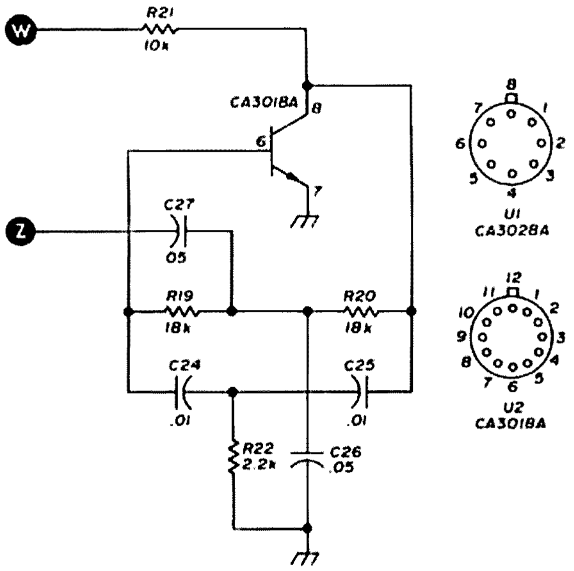
The conventional direct-conversion receiver of **fig.** 3 uses an RCA CA3028A as the product detector and Q1 as the oscillator. Inclusion of L7 on the L6 core eliminates the need for the rf choke normally found between pins 1 and 5 of U1. The entire audio section is contained in a single RCA CA3018A linear IC. The audio preamplifier uses one of the two Darlington-connected transistors; the audio amplifier uses a separate transistor on the same IC chip. Recovered audio is more than sufficient to drive high-impedance headphones. The 0.5-μF capacitor from the top of the audio gain control serves to attenuate some of the higher audio frequencies. Although this capacitor also reduces the overall audio output available, sufficient margin is left to run the audio gain control about one-third open for most operations.

The remaining transistor on the chip of U2 is connected in a twin-tee configuration and acts as the sidetone oscillator. Power is derived from the keyed +12 volts to Q2. The tone can be adjusted by varying R22. The volume may be more than required; if so, decrease the value of the sidetone oscillator output coupling capacitor C27 until a comfortable level is reached. To ensure that the audio preamp is firmly "off" during transmission, R8 upsets the base bias enough to saturate the transistor, thus preventing any signal other than the sidetone oscillator to be introduced into the audio amplifier.

cabinet layout

An LMB 101 Minibox (12 cubic inches

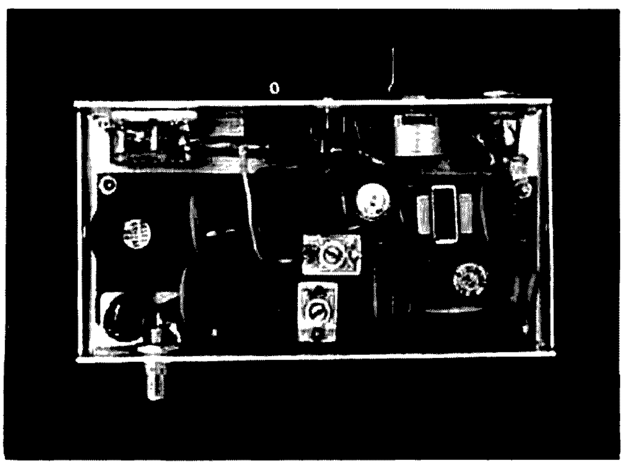




C11	3-35 pF trimmer (Arco 403)
J4	3.5-mm phone jack
L5	60 turns no. 28 enamelled on Amidon T-50-2 toroid core
L6	5 turns no. 22 enamelled wound around L5
L7	10 turns no. 22 enamelled wound around L5
R14	miniature 10k pot, audio taper
T1	miniature transformer, 10k primary, 2k center-tapped secondary (Calrad CR-75 or Calectro D1-722)

fig. 3. Schematic diagram of the receiver circuit.

or 197 cc) was selected as about the smallest available enclosure which provides enough panel space for the necessary controls. Elimination of the vxo feature would have permitted a smaller enclosure, but it was felt that the flexibility provided by the vxo was essential. An internal key, a simple pushbutton switch, was included on the cabinet top. This made upside-down mounting of the PC board necessary. The external key jack may be eliminated if desired.



Miniature 7-MHz transceiver is built into small, 12-cubic-inch LMB 101 Minibox.

printed-circuit board

As can be seen from the photographs, more than enough space for the board is available; component density is fairly compact, but not unmanageably so.* The layout of components on the board was determined after considering the location of PC board inputs and outputs which give minimum interconnection lengths to the panel controls. By using an elevated

*Printed-circuit boards for the miniature 7-MHz transceiver are available from MFJ Enterprises, Post Office Box 494A, Mississippi State, Mississippi 39762 for $3.75, post-paid.

heatsink on Q2, it is possible to locate some components under the heatsink near the body of Q2. Toroid coils are mounted vertically to conserve space, and held in place with Q-dope after soldering. The ¼-watt resistors are also vertically mounted throughout. U1 and U2 are wire the PC board and install it. Four small bolts serve as corner mounting posts with washers and nuts providing the appropriate spacing between the board and chassis. Dry transfer labels were applied and then given a light coating of acrylic spray to prevent rubbing off.

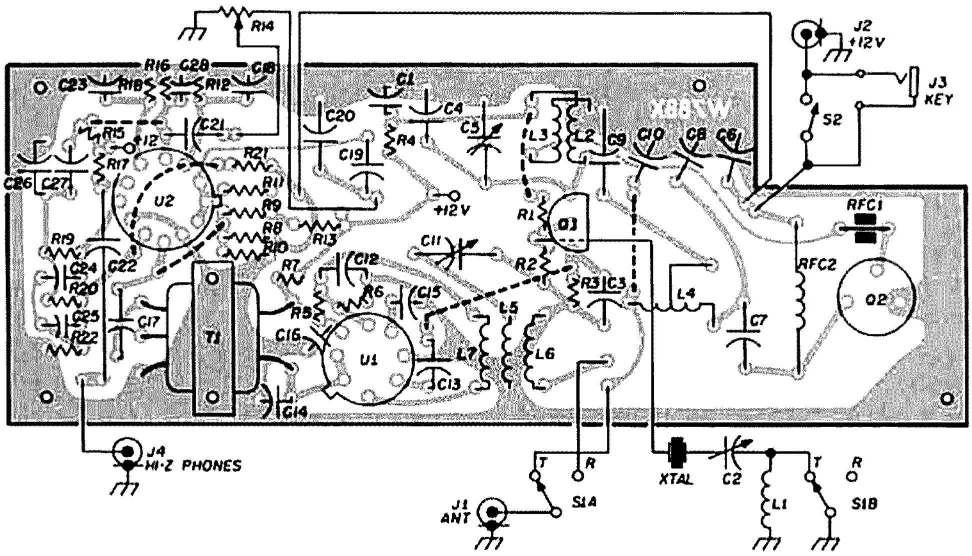


fig. 4. Component layout for the miniature 7-MHz transceiver. Printed circuit is shown in fig. 5.

mounted in IC sockets which are press-fit into appropriate-sized holes, and held in place by epoxy on the underside and the component leads soldered to the socket pins. Check the value of each of the ¼-watt resistors used with a reliable ohmmeter; the tolerances indicated are often exceeded.

construction

It is helpful to first mount all chassis controls. Size and layout of the PC board paper template can then be verified before actual construction and etching of the board. After the board is etched and drilled, drill the four corner post-mounting holes in the chassis top. Then

Alignment requires only a separate 7-MHz receiver and a dc milliammeter or vom capable of reading 250 mA. With the antenna connected and S1 in the *receive* position, peak C11 for maximum received signal; this completes the receiver alignment.

Connect a 51-ohm, 1-watt resistor to J1 and place a milliammeter in series with the +12-volt supply to J2. Place S1 to *transmit* and depress the key. Adjust C5 for maximum meter reading and good keying characteristics in the monitor receiver. Approximately 200 mA should be indicated; remember that receiver current drain and oscillator current are included in the metered current.

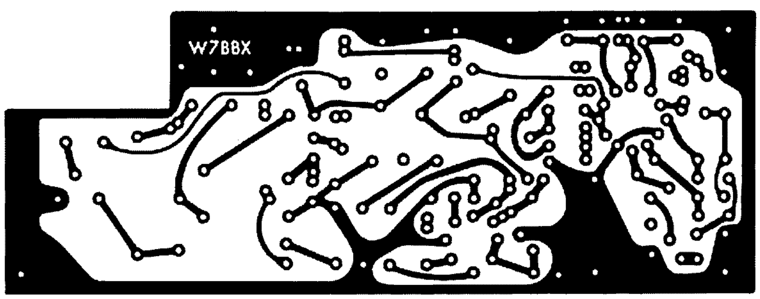
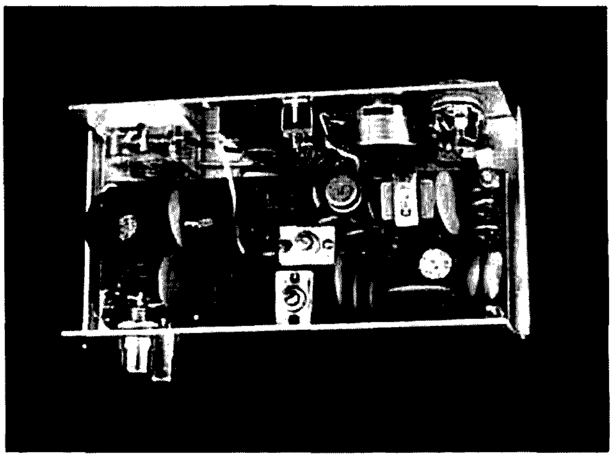


fig. 5. Full-size printed circuit board for the 7-MHz transceiver. Component layout is shown in fig. 4.

To adjust the frequency offset required during transmit, monitor the oscillator frequency in the *receive* position and transmitter output frequency in the transmit position when keyed. Adjust the number of turns on L1 to make the two coincide. Although 7 μH



Component layout for the 7-MHz transceiver. In this photograph the transmitter output transistor is to the left, the receiver input circuit to the right.

was required in my version to provide the necessary 1-kHz offset, this value may vary. L1 can then be cemented to the rear of the crystal socket with Q-dope.

Reconnect the antenna to J1 and try not to appear surprised when you consistently get 569-599 reports from single-hop contacts. Experience has shown, however, that double-hop QSOs from 2000 to 3000 miles are generally in the 339-559 range with a well-matched dipole up 18 feet (5.5 meters).

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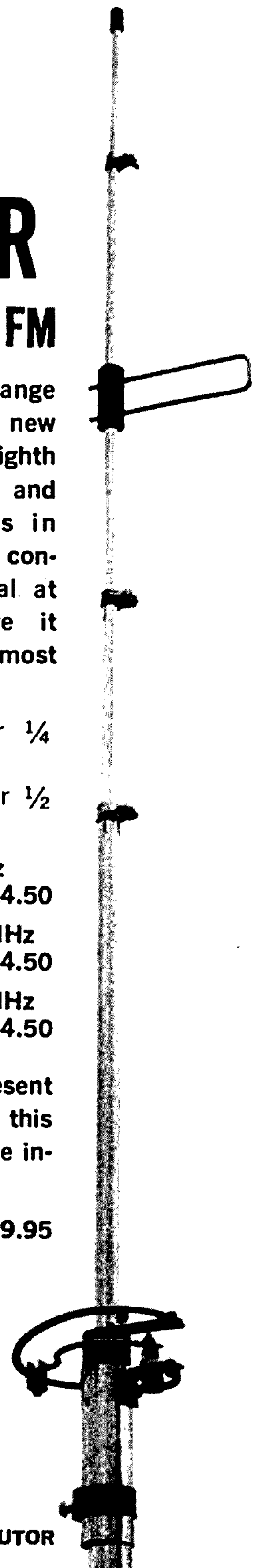
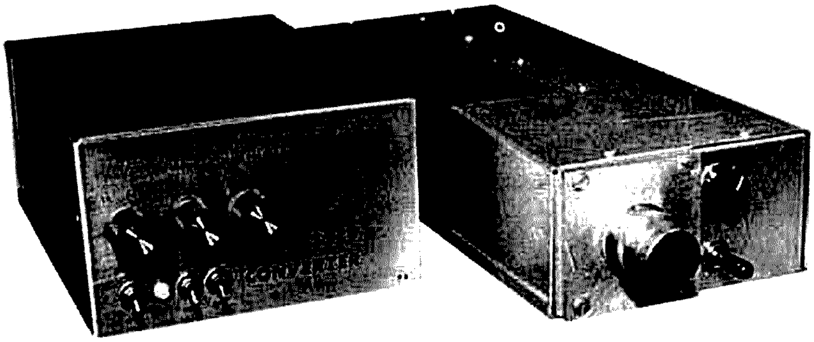
3. Wes Hayward, W7ZOI, "The Micromountaineer," *QST*, August, 1973, page 11.
4. *RCA Linear Integrated Circuits*, Technical Series IC-42, 1970, page 312.
5. *RCA Transistor Thyristor and Diode Manual*, Technical Series SC-15, 1971, page 547.

ham radio





fast-scan camera converter for sstv

Complete construction details for an economical approach to sstv picture generation using a surplus FS camera

Gordon P. Howlett, WA9UHV, Kokomo, Indiana 46901

Conversion of surplus fast-scan (FS) closed-circuit TV cameras to slow-scan television (sstv) standards has attracted many hams to the exciting field of picture transmission on the hf bands. W9NTP[1] and others have pioneered the sampling conversion approach to sstv picture generation. W3EFG[2] introduced an sstv sampling converter, using discrete components, that produces high-quality pictures via camera conversion.

This article presents an sstv converter using some of the low-cost, high-performance integrated circuits that have become readily available. Also included are several FS camera conversion techniques that may be helpful to interested experimenters.

The project began with the following objectives:

1. High-quality sstv pictures with minimum cost.

2. Use of simple, proven circuits and

readily available ICs were feasible and advantageous.

3. Derivation of all timing signals from the 60-Hz power line.

4. Features such as video reversal and fractional scan.

5. Stability with temperature.

6. Simplified FS camera conversion.

preliminary steps

1. After studying the FS camera schematic, locate and identify the various circuits in the camera, including polarities and absolute levels of signals; this information will speed the camera/converter interface task.

2. Using an oscilloscope, determine the high and low sides of the vertical deflection yoke coils and record the amplitude of the 60-Hz ramp at the yoke.

3. Locate the sync insertion circuit; this will usually be a single transistor whose base is driven with signals from both the vertical and horizontal deflection amplifiers. Generally this transistor will be located at or near the output of the video amplifier.

4. Observe and record the polarity and level of the vertical sync pulse; this information will be helpful in determining the type of interface (if any) required between the converter sync amplifier and camera.

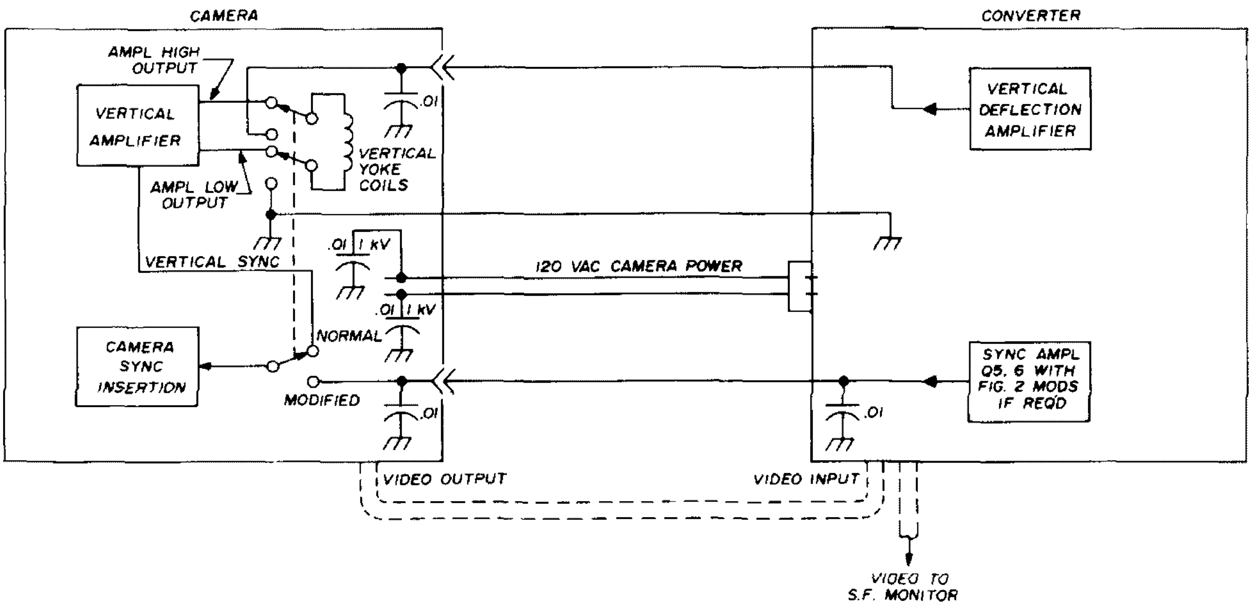


fig. 1. Typical camera-to-converter connections. All leads out of the camera (except video output) should be bypassed. Shielded cable will help with RFI reduction.

camera modifications

The frame rate of the camera must be reduced from 60 to 15 Hz. In order to overcome the difficulties in obtaining a linear 15-Hz sweep voltage from the camera's 60-Hz vertical deflection amplifier, I included a simple deflection amplifier with the converter. This amplifier is similar to one used by W3EFG. Also, to ensure that the vertical size of the FS picture is identical in both the FS and SS modes, I decided to switch size controls rather than vertical ramp capacitors. This method eliminates ramp capacitor selection for the SS mode.

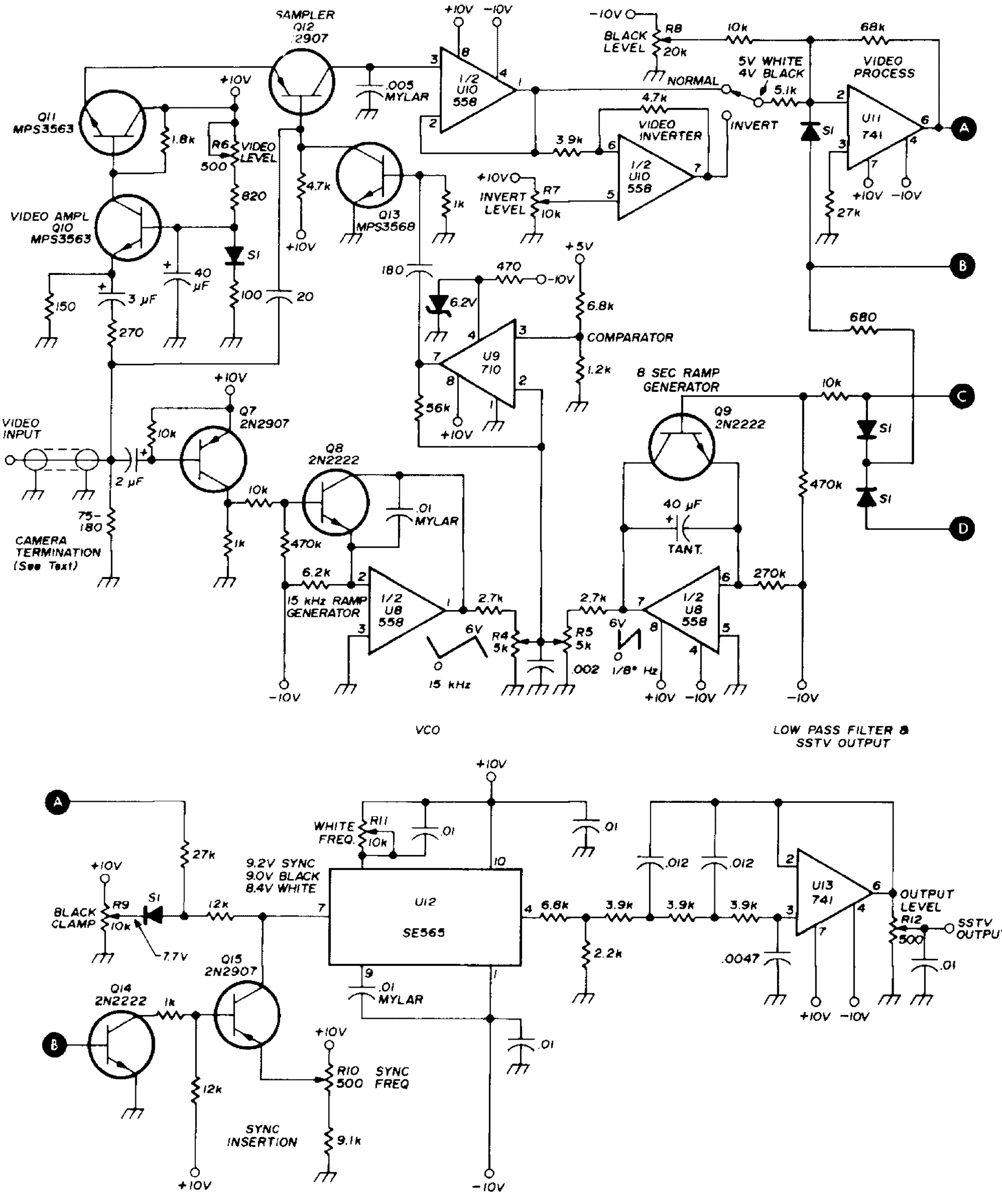
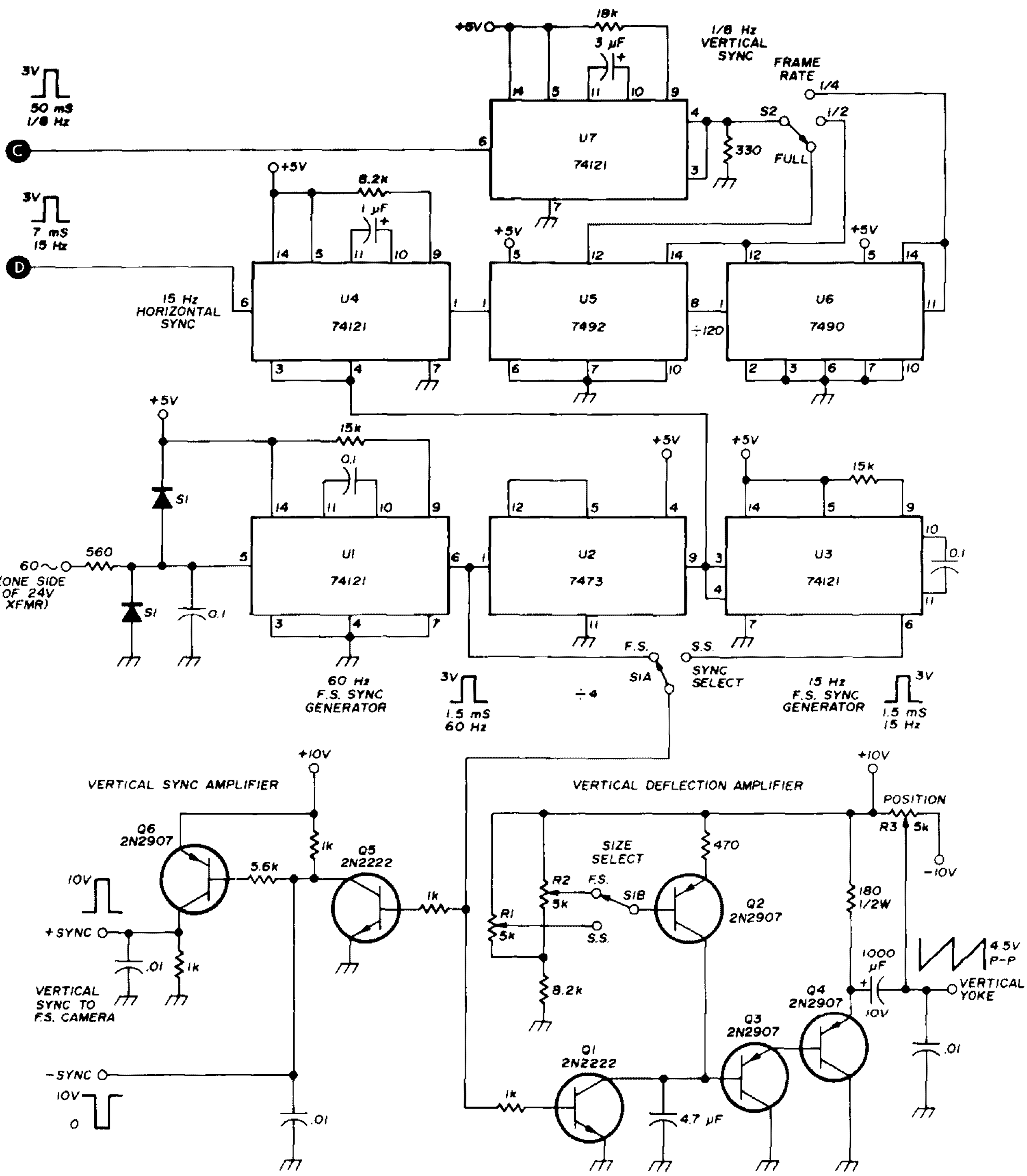


fig. 2. Converter schematic. Voltages and waveforms are for reference only; see text.

A typical camera/converter interface is shown in **fig. 1**. Vertical yoke drive and vertical sync information is supplied to the camera from the converter. A switch may be mounted on the camera case to select between converter-supplied signals or normal camera operation. If the camera will not be used in other non-sstv applications, the switch may be omitted and direct connections made to the camera.

interface circuits

The sync amplifier, Q5 and Q6 (**fig. 2**), produces positive and negative sync pulses between zero and 10 volts. Many



cameras use a negative power supply voltage. In this case, you may need to incorporate one of the interface circuits in **fig. 3**, since quite likely the sync pulses in the camera will swing between zero volt and some negative potential. As mentioned earlier, careful study with an oscilloscope before beginning camera modification will show you which combination of sync amplifiers in **fig. 2** and interface circuits in **fig. 3** are required.

As with the W3EFG converter, the camera (or yoke assembly) is rotated 90 degrees to lie on its left side so that the top of the picture will be at the viewer's right side on the FS monitor.

camera power supply

The camera's primary power is supplied from a connector on the converter. This connector is wired so that the camera cannot be operated without the

converter turned on and the deflection amplifier active. Connections in this manner help prevent serious vidicon target burn resulting from loss of scan.

Stray magnetic fields may occur from a camera-mounted power supply. These fields can cause distortion on the FS picture, which appears as wavy kinks or vertical bars on the sstv picture. Quite possibly mu-metal shielding around the power transformer and/or vidicon, or a piece of thin copper strap formed around one leg of the power transformer core as a shorted turn, will suppress the distortion.

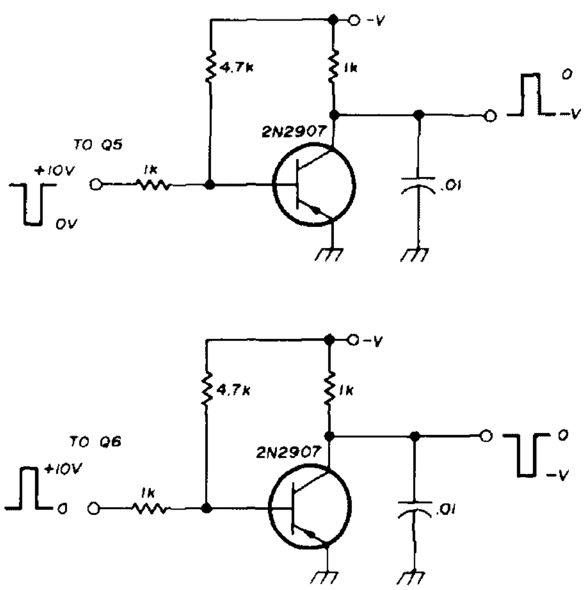


fig. 3. Special interface circuits for cameras with negative sync signal requirements.

If these measures don't succeed, the power transformer may be remotely mounted in the converter package as mentioned later in the construction section.

picture contrast

Some experimentation with the camera video termination may yield higher contrast on sstv. With my converted FS cameras, I found that terminations higher than the standard 75-ohm value will yield usable pictures at lower light levels. Fig. 2 indicates that the termination may be between 75 and 180 ohms. Once the complete system is operating, the termination should be optimized.

video monitor

I made a simple modification to my old TV set so it could serve as a video monitor. As shown in fig. 4, a 30-μF capacitor is connected to the output of the video filter. A short piece of coax (less than 1-foot [30.5 cm] long) is run to a connector on the chassis. Plug the camera video output into this coax, through an additional length of cable, add a camera termination, and you have a picture. The termination is not required if the converter (with its own termination) is in the circuit.

circuit description

Explanation of the sampling principle of scan conversion has been well covered elsewhere.[1] It is recommended that W9NTP's excellent articles be studied by those not familiar with the sampling technique.

The converter schematic and block diagram are shown in figs. 2 and 5 respectively. Line frequency (60 Hz) is applied to U1, producing 1.5 ms, 60-Hz pulses. These pulses are divided by 4 in U2 and shaped into 1.5 ms, 15-Hz pulses by U3. Switch S1A selects either 60- or 15-Hz pulses for the vertical deflection and sync amplifiers. Transistors Q1 through Q4 comprise a direct-coupled yoke driver similar to the one used by W3EFG.[2] Switch S1B selects the vertical ramp size for the FS and SS modes. A variable dc bias on the yoke centers the scan on the vidicon target. Transistors Q5 and Q6 interface the sync generators with the FS camera. Transistor Q6 may be omitted if negative sync is required by the FS camera, as mentioned earlier.

The sstv horizontal sync pulse width is determined by U4. A separate monostable multivibrator is used, since it's desirable for the 15-Hz FS vertical sync pulse (determined by U3) to be considerably shorter than the sstv horizontal sync pulse, thereby eliminating shading on the left side of the sstv picture.

The 15-Hz sstv horizontal sync pulses are used by ICs U5 and U6 to produce

the sstv vertical sync pulse. These ICs are connected so that switch S2 can select divide-by-30, divide-by-60 and divide-by-120 to produce ¼, ½, and full sstv frame rates respectively. The sstv vertical sync pulse width is determined by U7.

Sync stripper Q7 removes the 15,750-Hz FS horizontal sync pulses from the composite video. These pulses are used to produce a 15,750-Hz ramp by Q8 and one-half of U8. One-eighth Hz ramps are produced by Q9 and one-half of U8. The amplitudes of these ramps are adjusted by controls R4 and R5 to set the start (R4) and finish (R5) of the FS frame sampling. The summation of these two ramps is compared with the reference voltage on pin 3 of comparator U9, producing a sliding pulse at its output, pin 7.

Video amplifier Q10 and emitter follower Q11 amplify and shift the FS video to a dc level determined by the setting of the video level control, R6. The diode in the base bias circuit of Q10 is used for temperature compensation.

The sliding pulse from U9 is differentiated by the 1k, 180-pF RC combination, producing a voltage spike that switches Q13 and Q12 to sample the FS video. The 0.005-μF capacitor holds the sampled voltage until the next sample.

One-half of U10 provides an impedance transformation between the holding capacitor and succeeding circuits. The other half of U10 provides the video inversion feature. Control R7 sets a reference level such that the average video voltage selected by S3 is the same for either the normal or inversion function.

Video-processor IC U11 amplifies the sstv video and establishes a reference level, controlled by R8, which sets the range of the panel-mounted video-level control. Composite sstv sync is applied to U11 to blank the video during sync insertion. The black clamper, R9, prevents video excursion below 1500 Hz.

Composite sstv sync, applied to transistors Q14 and Q15, clamps the vco voltage to the level set by the sync frequency control R10.

The vco, in an SE-565 phase-locked loop, is used because of its excellent linearity. Potentiometer R11 sets the white frequency. The square-wave vco output is filtered by a simple active low-pass filter, providing a low source

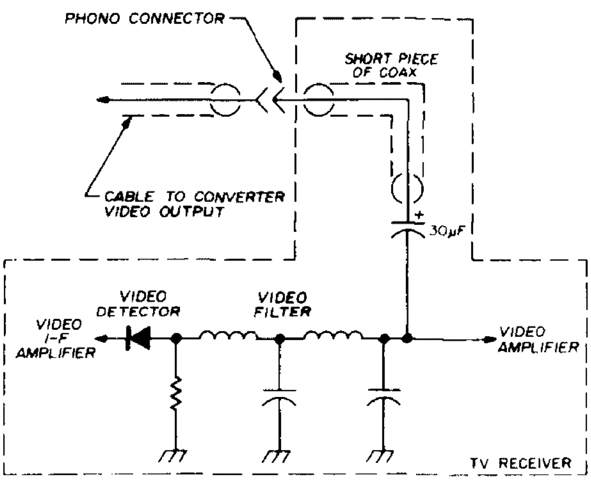


fig. 4. Tv receiver modification for use as a video monitor.

impedance, through the output level control, R12, for driving external equipment.

The power supplies use integrated circuits to control series-pass transistors. The negative supply tracks the positive supply, resulting in a single adjustment. Five volts are obtained from a single IC. All supplies are current limited.

construction

Lead length and component placement are not critical in the sstv converter. My breadboard model has long leads and haphazard stage and component placement. The second model, shown in the photographs, was built using the technique I've found convenient for camera and monitor construction.

After board layout and approximate stage placement has been selected, B+ and ground busses are run on the vector board. Components are then mounted between the busses using short, direct connections. The result is a reasonably

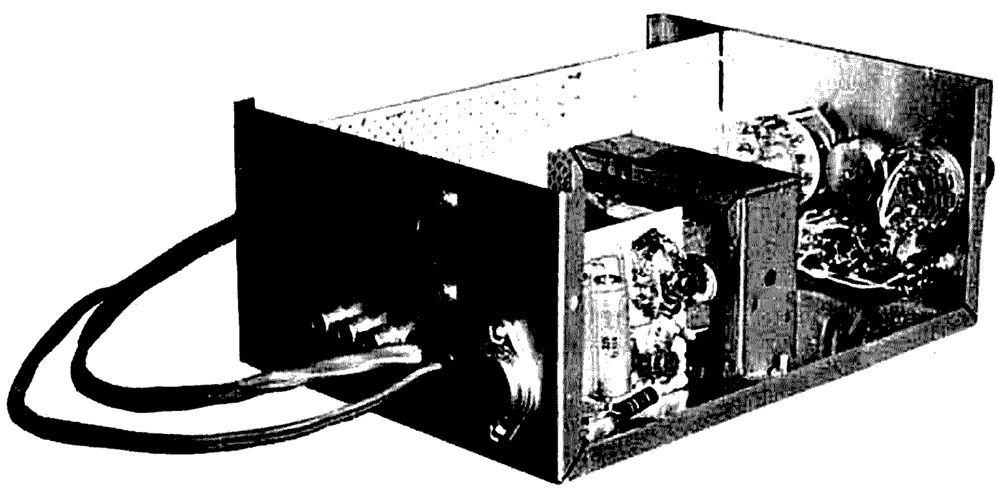
neat, one-sided layout that's easy to troubleshoot and later modify. Sockets are not used for the active devices because of the extra space and cost required, although they may be included if desired. Vector board pins are used only at board/wire interfaces. Teflon wire is recommended for board jumpers to prevent insulation burning during soldering.

Board size of the converter is 7½ inches by 3¼ inches (19.1 x 8.25 cm), which also includes a gray-scale generator not described here. The power supplies are located on a separate board near the rear. The cabinet is a 8 x 6 x 3½ inch (20.3 x 15.2 x 8.9 cm) Minibox.

If it's desired to build the camera and converter as a single unit, removal of the camera's power transformer should provide adequate room for the converter. Since layout is not critical, the converter can be packaged on two or three small boards to suit the available space. Camera and converter power transformers may be mounted on a separate chassis and connected to the camera through a multiconductor cable. Radio-frequency interference difficulties may be reduced by single-unit packaging.

setup procedures

Disconnect the power supplies from the converter and set the ±10-volt adjustment. Check for 5 volts at U16. Reconnect the supplies and use an oscilloscope to trace the waveforms, shown on the schematic, through U1, U2, U3 and the vertical deflection amplifier. With the scope at the vertical yoke connection, set S1 to FS mode and adjust R2 for a maximum, linear ramp signal without positive peak clipping. Switch S1 to SS mode and adjust R1 as above. Set R3 to center rotation.



Left-rear of converter. The power transformer is mounted in the center for balance.

Turn the converter off and connect the camera to the converter as shown in fig. 1. If any of the special interface circuits are required, they should also be in the circuit. With power applied, a locked raster should appear on the FS monitor. Be sure S1 is in the FS position. With the aid of a test pattern or a circle on a sheet of paper, adjust R2 and R3 for a centered, symmetrical picture. Note the amplitude of the 60-Hz ramp on the vertical yoke with the scope, switch S1 to SS, and adjust R1 to the same amplitude. If difficulty is experienced in locking the

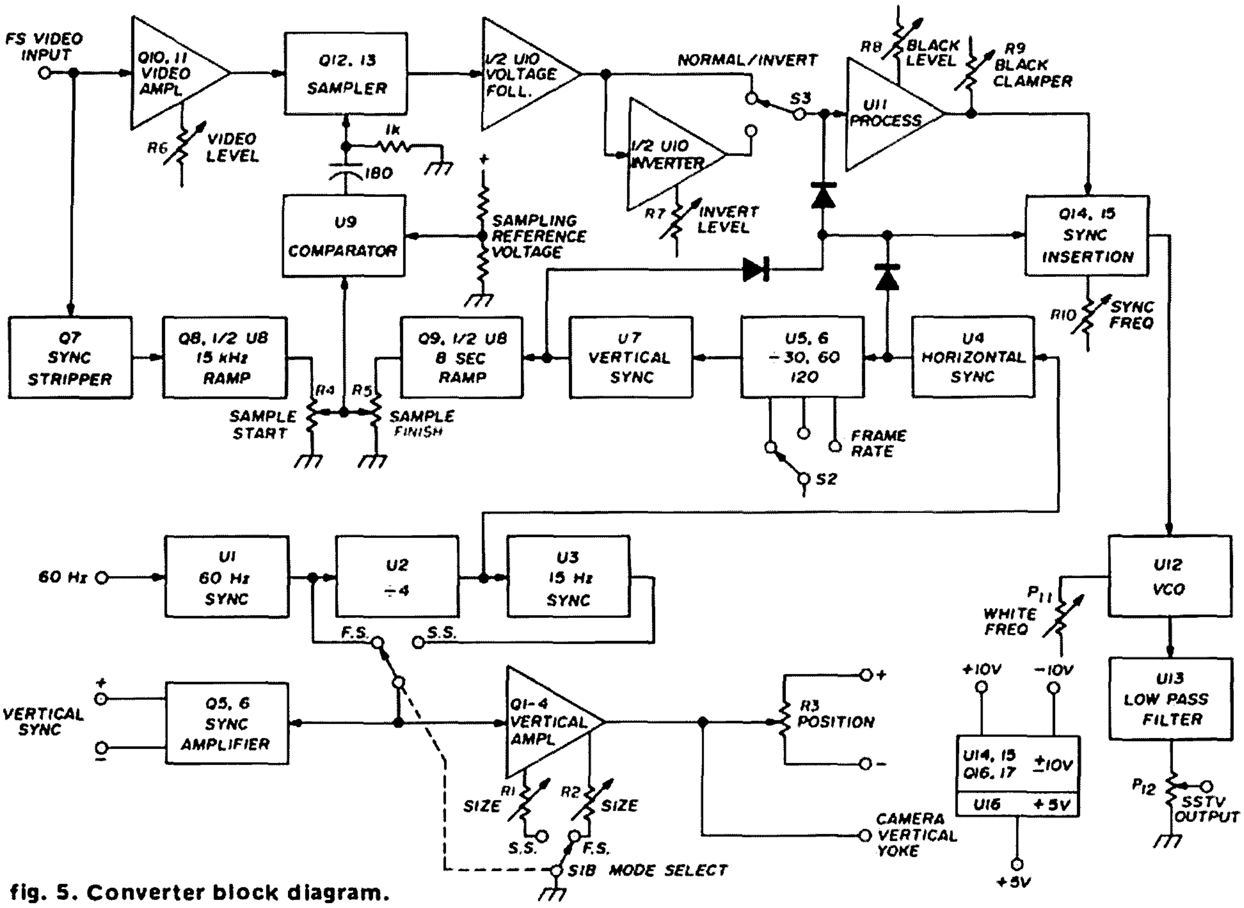


fig. 5. Converter block diagram.

raster on the monitor, experimentation with the sync insertion connection may be required.

Using the scope, check the following points for the waveforms shown on the schematic:

1. U4, pin 6: 15-Hz pulses.

2. U7, pin 6: 1/8-, ¼-, and ½-Hz ramps, depending on S2 position.

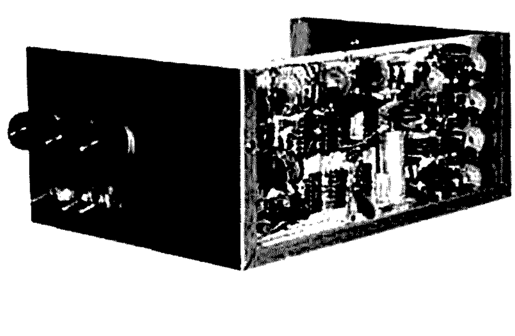
3. U8, pin 7: 1/8-, ¼-, and ½-Hz ramps, depending on S2 position.

4. U8, pin 1: 15,750-Hz ramps.

The output of the comparator, U9, pin 7, is a sliding pulse that changes width during picture scan. Adjustment of R4, scan start, and R5, scan end, will produce a continuous sliding pulse that changes width smoothly and without hesitation. The differentiated spike will be visible as a thin black line on the FS monitor, moving from right to left. Its position on the screen indicates the part of the FS picture being sampled.

Sstv has a 1:1 aspect ratio (FS is 3 high to 4 wide), so a narrow portion of the left- and right-hand sides of the FS picture will be lost. Controls R4 and R5 have some interaction, so patience is required to center the sampled portion of the FS picture.

The setup of the vco and clamper should be made in a step-by-step procedure initially. The FS camera (with its lens capped) must be connected to the converter. Preadjust the following controls as indicated:



All circuitry (excluding power supply) is mounted on Vector board along one side of the box.

1. R9 to 10 volt end.

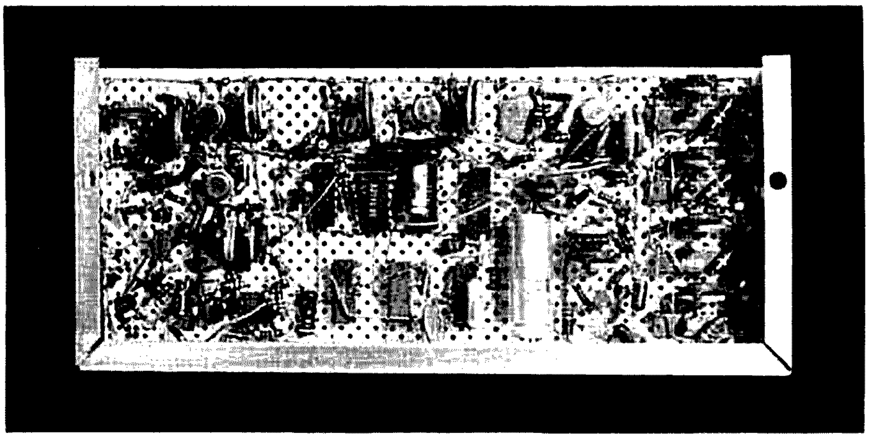
2. R6 to maximum resistance.

3. R11 to midrange.

Proceed with final adjustment as follows:

1. Adjust R8 from ground until the vco frequency (as monitored at the sstv output jack) begins to decrease.

2. Set R11 to 2300 Hz.



Main converter board. Circuits in the left quarter are the video amp, ramp generators, comparator, sampler, and video processor. Lower center contains the counters and sync pulse generators. Upper center is the vco and output filter. Immediately to the right of the large cap is the yoke driver and sync amp. Circuitry in the right quarter is the grey-scale generator.

3. Set R6 to minimum resistance.

4. Adjust R9 for 1500 Hz.

If it is not possible to adjust the vco frequency over the full 1500- to 2300-Hz sstv video range with the video level control, readjust the black level control, R8, and repeat steps 2 through 4. With a 1k resistor connected between the base of Q14 and 5 volts, set the sync frequency control, R10, to 1200 Hz. Set R6 to 1900 Hz, switch S3 to invert, and adjust R7 to 1900 Hz. Set S3 to normal.

The voltages and waveforms were measured in one of my converters; they are for reference only and may vary considerably between units.

With the lens uncapped, adjustment of the video level control and lens opening will produce a sstv picture.

component substitutions

I've built two converters using identical component values and observed identical results. It's possible, however, to change some component values to suit personal preference. The 68k resistor at U11 may be revised to change the sstv contrast, or a control may be mounted on the panel if desired. Likewise, the contrast in the invert mode may be adjusted by changing the 4.7k resistor at U10. Sstv sync pulse width may be changed by adjusting the timing resistors in U4 and U7.

With the component values shown, the video circuits will operate at high gain. It is then possible to obtain useful sstv pictures using less light and/or lower lens openings. In either case, camera setup is simplified.

device selection

Transistors Q10, Q11 and Q13 are high-frequency devices; Q12 is a low-leakage, high-frequency unit. All other transistors except Q16 and Q17 are small-signal devices. You may substitute providing you use silicon types. ICs U8 and U10 are dual op amps, which may be single 741s if desired. Diodes, except those in the power supply, are small-signal devices of the 1N914 variety.

operation

Only two adjustments (assuming the lens is focused) are required to set up the camera/converter for pictures. I use the lens opening, or amount of light, to control the contrast and set the video level control so the video swings from full black to full white with plenty of gray scale in between. Be sure your monitor's brightness and contrast controls have

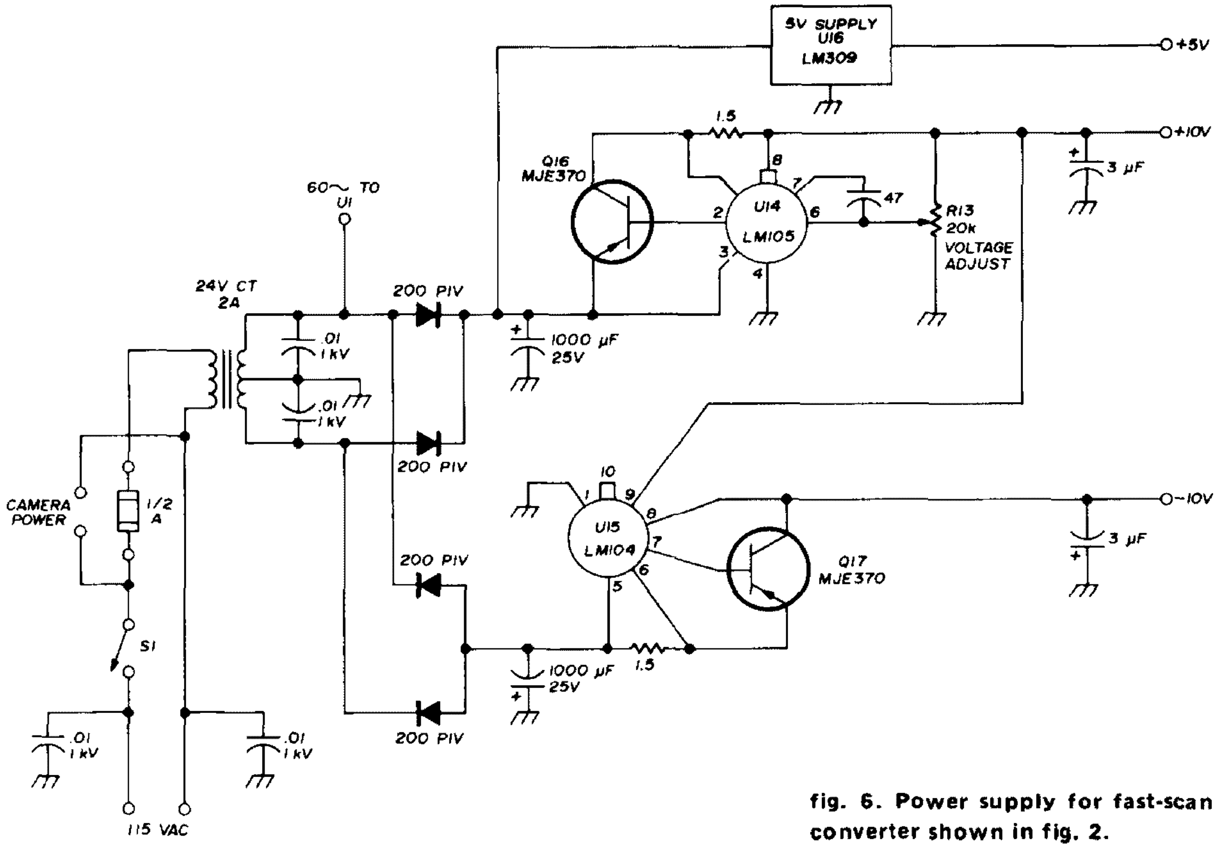


fig. 6. Power supply for fast-scan converter shown in fig. 2.

been set using a gray-scale tape or generator, as the monitor is your reference for camera adjustment.

I use a 100-watt bulb in a gooseneck lamp and a f/16 lens opening for photographs. For lettering, I open the lens to f/11 so the video will swing from full white to full black with no gray scale. Live pictures are taken with normal room lighting and a f/4 or f/5.6 lens setting.

summary

The conversion of a used, closed-circuit FS camera to slow-scan provides an economical approach to sstv picture generation. Furthermore, an old black and white TV may be pressed into service to allow fast, accurate focus and lighting adjustments.

The scan converter described here produces high-quality sstv pictures using inexpensive integrated circuits to reduce complexity and simplify construction. With the addition of only a few components, fractional frame rate and video reversal features are included.

I've attempted to use proven circuitry where possible and not try to reinvent the wheel. Several circuits, created by others, have been used since they provided the best performance with the least complexity and cost, which was one of my goals. I would be interested in hearing from readers who build this converter. Those interested in using this basic converter with a FS camera previously modified for use with the W3EFG converter (divide-by-4 and sync/deflection amplifiers not required) are invited to write for interface details. Please include a self-addressed, stamped envelope.

acknowledgements

I would like to thank James Mathews, W9KYS, for assistance with the photographs and K.O. Learner, K9PVW, for stimulating discussions during the course of the work.

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2. R.F. Stone, W3EFG, and A.B. Schechner, W3YZC, "Conversion From Fast-Scan to Slow-Scan Television," *ham radio*, July, 1971.

ham radio

a versatile autopatch system for vhf fm repeaters

Complete design information for an autopatch system that includes access control and protective features

R.B. Shreve, W8GRG, 2842 Winthrop Road, Cleveland, Ohio 44120

An autopatch for fm repeater use is not just an automated phone patch. A good autopatch should have protective features not necessary on a phone patch at an attended station. On the other hand, certain features that I consider desirable in a phone patch add nothing to an autopatch installation.

Both phone patch and autopatch should provide undistorted audio coupling from the telephone line to the transmitter and from the receiver to the telephone line. Freedom from distortion between the receiver and telephone line is particularly important in an autopatch working into a Touch-Tone central office. Provision should be made on both phone patch and autopatch to set the audio gain at levels acceptable to the telephone system, and to provide a desirable amount of modulation to the transmitter.

phone-patch requirements

Beyond acceptable audio, the two systems have little in common. Features desirable in a phone patch to be used at an ssb station have been described in another article[1] and include facilities for three-way conversation, enabling the operator to talk to both parties—the person on the telephone as well as the one on the radio—using the station microphone and monitoring with the station speaker or headphones. Vox operation of the transmitter by the party on the telephone is also an advantage. Successful accomplishment of these objectives can involve fairly sophisticated hybrid circuitry and incorporation of a number of amplifiers for level control and circuit isolation.

autopatch features

An autopatch system does not require vox operation or isolation of the transmitter audio input from the receiver output. Monitoring is the responsibility of the control operator, usually at a remote control point. The autopatch circuit is essentially one for two-way communication, with switching between transmit and receive controlled by the party originating the radio contact.

However, the autopatch system does need a control system which will permit the caller to gain access to the telephone line and originate and terminate a call. It should also have reasonable protection against accidental activation, and automatic time-out if the calling party neglects or is unable to terminate the call. It should also be possible for the control operator to interrupt a call if necessary, without turning off the repeater.

Optional control features include protection against toll calls, and a provision working successfully on a busy 146.16/76 repeater.

simple autopatch

The audio circuit for the simple autopatch is shown in **fig. 1**. This patch was constructed from a Healthkit HD-15 Hybrid Phone Patch, and uses most of the kit components and circuitry. Except for the Touch-Tone decoder and power supply the unit is completely contained in the 9¼ x 3½ x 2½-inch (23.5 x 8.9 x 6.4-cm) cabinet supplied with the kit.

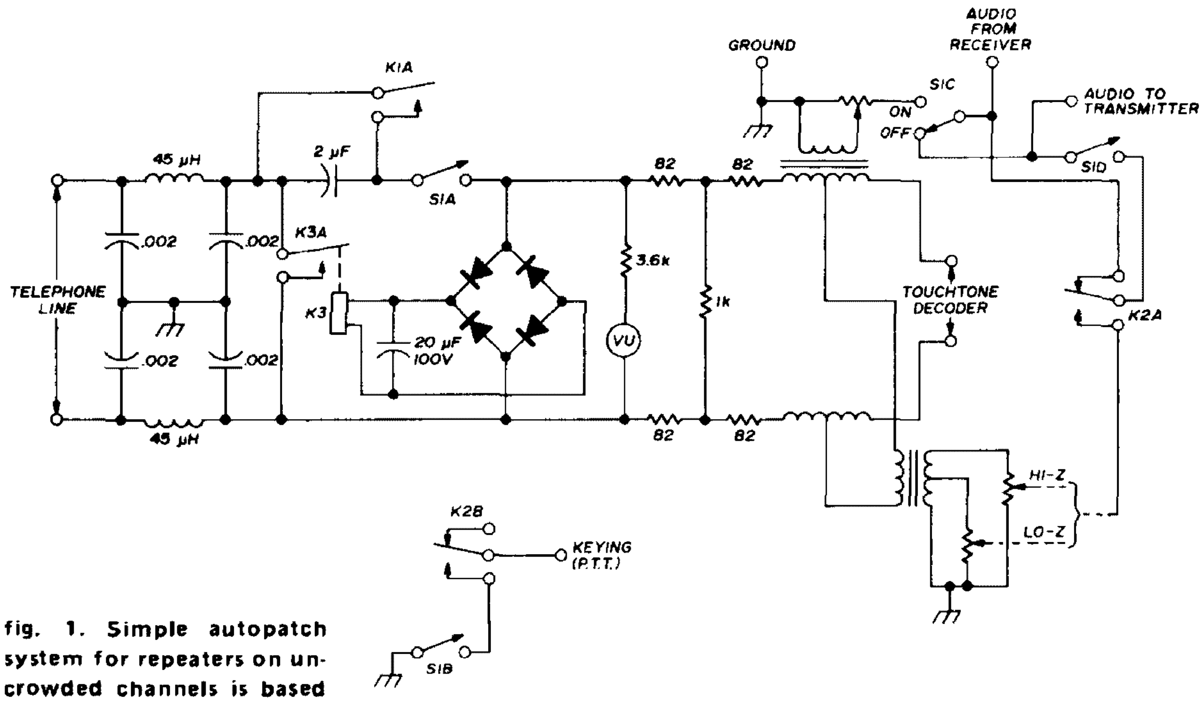


fig. 1. Simple autopatch system for repeaters on uncrowded channels is based on a Heathkit HD-15 hybrid phone patch kit. Control logic and timing circuits are shown in fig.2.

whereby persons knowing what number to call can dial the autopatch line, monitor the repeater input frequency, and if the control system permits, send control signals over the telephone line. Some groups have gone further, and made it possible to originate calls through the repeater from a telephone—sort of a "reverse autopatch." Restricting this capability to licensed operators is a problem those choosing to install such a system must solve for themselves.

This article describes two autopatch systems that I have built. One is comparatively simple, suitable for use on a repeater on an uncrowded channel. The other, with more protective features, is

The Heathkit HD-15 phone patch includes a hybrid circuit which isolates the receiver audio output from the transmitter. As previously noted, this isolation is unnecessary in an autopatch. However, the hybrid transformer provides excellent quality audio coupling between receiver and telephone line, and from line to transmitter. The null adjustment balancing network is replaced by the Touch-Tone decoder which operates the phone patch control circuits. The control circuits combine solid-state logic with relay switching.

Operation of the patch is as follows. Switch S1 is a manual *on/off* switch, which in the *off* position disconnects the

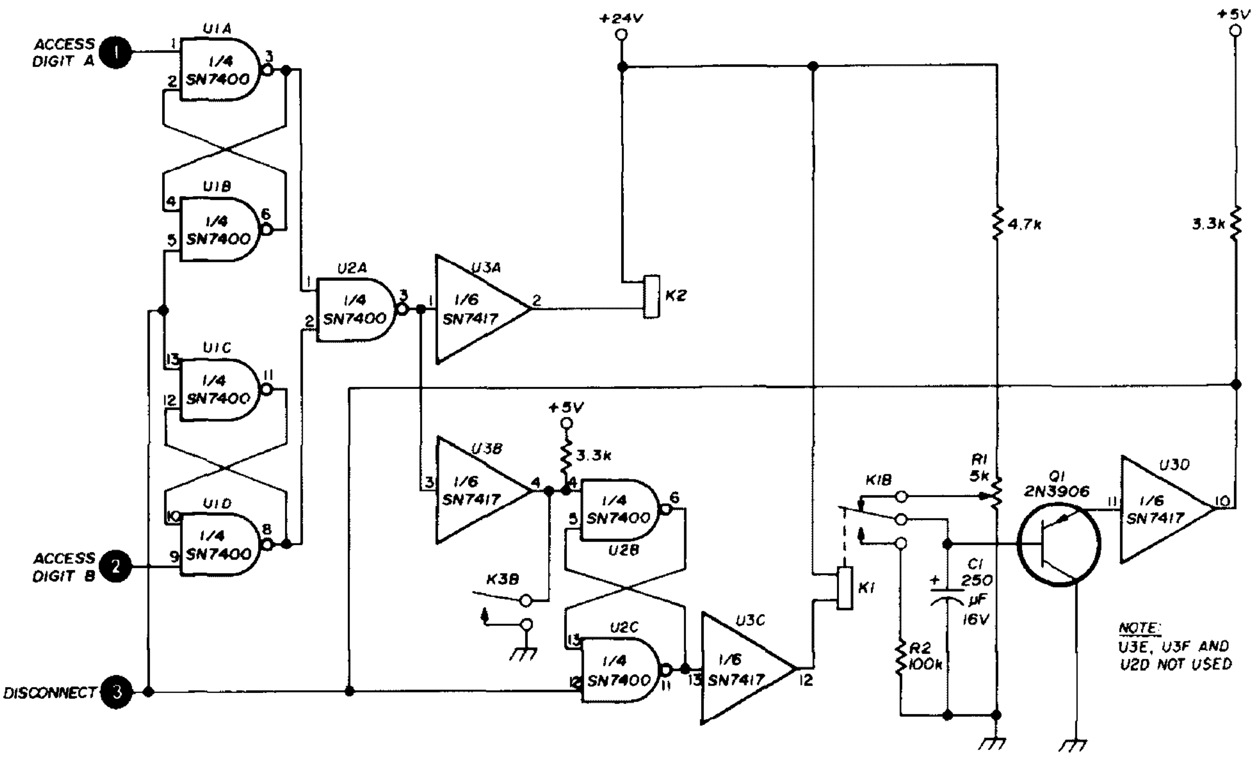


fig. 2. Simple control logic and timing circuit for the autopatch system shown in fig. 1.

phone patch from the repeater and telephone line, and bypasses the repeater *audio signal from* receiver to *transmitter*. This switch is normally left *on* to permit autopatch operation.

When a call is to be made, relays K1 and K2 are operated by the Touch-Tone-controlled logic. Contact K2A opens the direct audio path between receiver and transmitter and connects the phone patch output to the transmitter input. Relay contact K2B keys the repeater transmitter. Contact K1A establishes a dc path for the telephone circuit, signalling the central office and bringing up the dial tone. Both these relays stay closed until the call is terminated. Everything said on the telephone and radio is transmitted over the repeater.

Relay K3 and the bridge rectifier compose an automatic telephone answering circuit. If the autopatch telephone is called by another party when manual switch S1 is *on*, rectified ringing current closes K3 momentarily, "answering" the telephone and stopping the ringing. Relay K1 closes, K2 does not. The calling party can monitor the repeater receiver over the telephone line, and transmit Touch-Tone commands to the Touch-Tone decoder to operate the patch or other repeater control functions. The connection should be terminated by the calling party by transmitting the disconnect signal before he hangs up his telephone.

The control logic shown in **fig.** 2 recognizes a two-digit access code and a single-digit disconnect signal. When the two access digits are received by the decoder it generates a logic zero, or ground, at inputs 1 and 2, operating relays K1 and K2 and activating the phone patch as described earlier. The disconnect signal is decoded as a logic zero at terminal 3, releasing both relays. While the disconnect signal will usually be transmitted by the calling party on completion of the call, it can be used by the control operator if necessary.

Automatic time limiting is provided with the circuit controlled by relay contact K1B. In the standby condition, capacitor C1 is charged to a level determined by the setting of potentiometer R1. When K1 is picked up, the capacitor is discharged through resistor R2. When

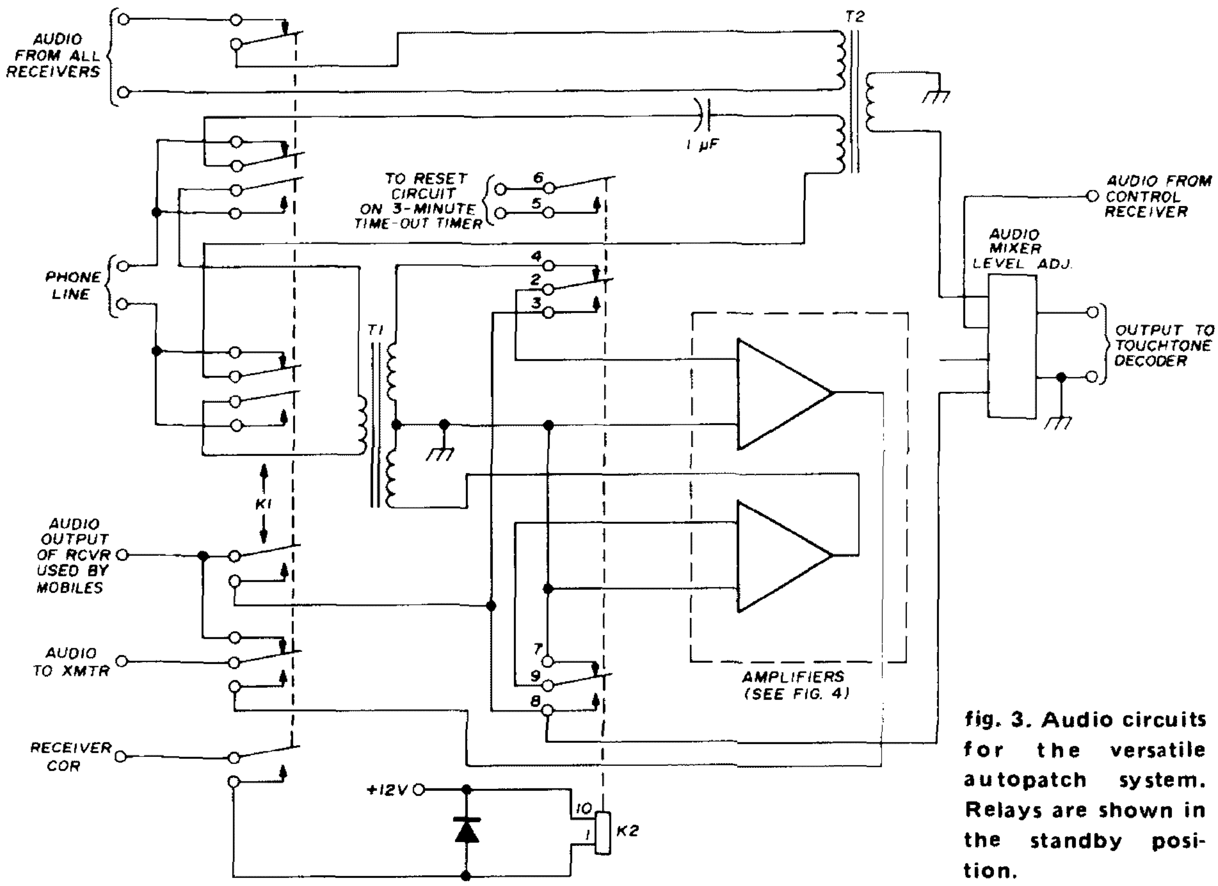


fig. 3. Audio circuits for the versatile autopatch system. Relays are shown in the standby position.

its charge falls to the level at which Q1 conducts, an automatic disconnect signal is generated at terminal 3. The timing cycle is also started by relay contact K3B, when relay K3 answers an incoming call. Of course, no one calling in on the telephone line to monitor the receiver would normally forget to transmit the disconnect signal—but K3B is still a good precaution, let's say, against someone calling a wrong number!

Performance of this simple autopatch is excellent as far as audio quality and circuit reliability are concerned. Its big disadvantage is lack of security—it can be activated by the two access tones in either AB or BA order, and there is no protection against long-distance calls. A lesser disadvantage is the lack of any way to boost the audio level from the telephone line to the repeater transmitter; it is sometimes hard for a mobile operator in heavy traffic to hear a soft-voiced woman when she answers the telephone. A more versatile autopatch system which overcomes these limitations is described below.

versatile autopatch

Although the autopatch system described previously provides excellent, reliable operation, it lacks some desirable security and operational refinements that enhance autopatch operation. The unit described here is more versatile, better protected, and incorporates audio amplifiers which *permit adjustment of audio* levels to the telephone line and transmitter. Construction is modular, so you can select the features you want and program the logic to fit your own requirements.

audio circuits

The audio circuits for the versatile autopatch system are shown in **fig. 3.** Component selection is not critical. Relay K1 is a multi-contact telephone type relay, and K2 is a 12-volt relay with one single-throw and two double-throw sets of contacts. Transformers T1 and T2 are audio transformers with three windings; any good-quality transformer with impedance of 400 to 1200 ohms and a 1:1 or 2:1 ratio should work. Surplus 400-cycle

power transformers with 117-volt primaries and 300-volt center-tapped secondaries have been used successfully.

The circuit diagram shows the relays in standby position. Audio from the receiver used by mobile stations goes directly to the transmitter. The telephone line is connected, through a blocking capacitor, to transformer T2. Another winding of T2 is connected to an input of the mixer which drives the Touch-Tone decoder. The third winding is connected to an audio source that monitors all the repeater inputs except the control receiver. This arrangement permits control operators to call the autopatch number on the set of contacts on relay K1, but keying it from the control logic permits more versatile operation, as explained later. Relay K2 is connected by a set of contacts on K1 to the mobile receiver COR which enables the mobile operator to switch the amplifier inputs, controlling what goes out on the telephone line and on the air. Note that this gives the mobile operator the ability to cut off any potentially objectionable remarks that might be made over the telephone by simply pressing his microphone button; he does not have to deactivate the phone patch.

The phone patch may be wired so that the repeater will repeat both sides of a

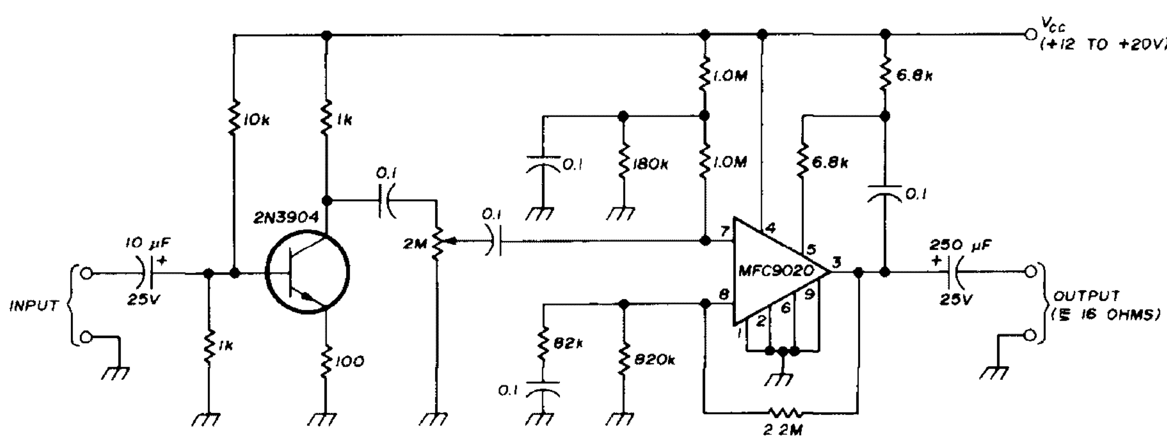


fig. 4. Audio amplifier circuit for the versatile autopatch system uses a Motorola MFC9020 audio power IC. Maximum power output is approximately two watts.

telephone, listen to the repeater inputs, and send instructions to the decoder over the telephone line. It also permits Touch-Tone signals from all receivers, such as those used to open a guarded input, to reach the decoder.

When the patch is activated audio from the mobile receiver is disconnected from the transmitter and connected to a phone patch amplifier and the Touch-Tone decoder. The telephone line is disconnected from T2 and connected to T1; the dc path through the transformer brings up the dial tone. The audio output that monitors all receivers is disconnected from the Touch-Tone decoder so only the mobile and control receivers can transmit instructions while the phone patch is activated.

The transmitter could be keyed by a conversation by interconnecting contacts 3 and 8 on relay K2 as shown in fig. 3. To have only the telephone party's side of the conversation repeated, omit the jumper between contacts 3 and 8, and connect 3 to ground. Details of the amplifier circuits are shown in fig. 4. They have tremendous gain and are capable of two watts output, which is what makes the impedance match of T1 relatively unimportant.

patch control circuits

The autopatch control logic, shown in fig. 5, is designed to operate with the solid-state repeater control logic described in an earlier article.[2] Activation of the phone patch requires two Touch-Tone digits in the proper sequence to set flip-flops U1 and U2. When U2 is set by

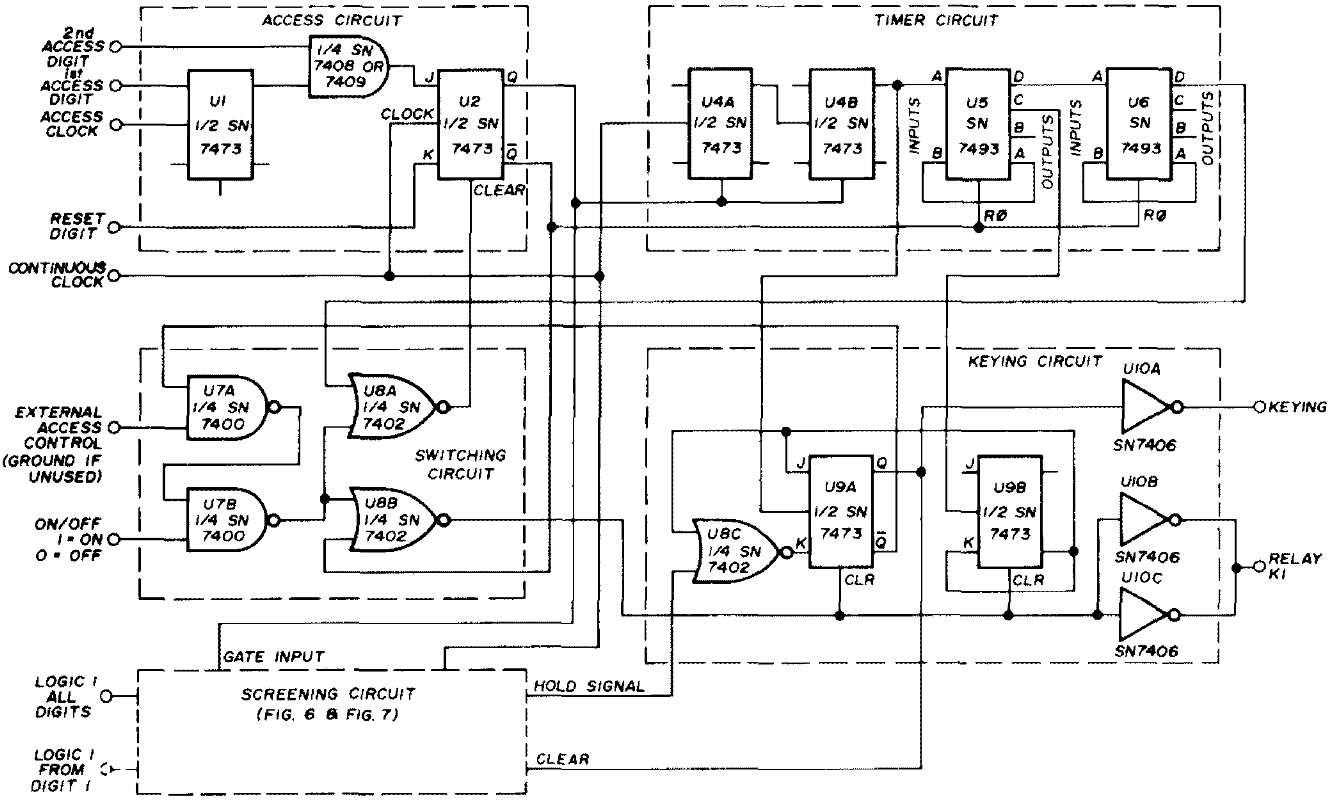


fig. 5. Control logic for the versatile autopatch system. Screening circuits are shown in figs. 6 and 7.

the final digit of the code, several things happen. The logic 1 at U2's Q output permits the timer to start counting clock pulses, and opens the gate to the digit screening module. At the same time, the logic zero at the $\overline{Q}$ output actuates the main patch relay K1 through the NOR gate U8B and the two inverter drivers, U10B and U10C, connected in parallel to handle the heavy relay current.

At the end of four clock pulses, the counter sets flip-flop U9A, keying the transmitter so that the dial tone is heard on the air. The caller then dials the desired number. The first digit of the number is checked by the screening circuit, and if it is acceptable a logic 1 is generated at the hold terminal, latching flip-flop U9A.

If no telephone number is dialed, or if the first digit is unacceptable to the screening circuit, no hold signal is generated and U9A is not latched. When flip-flop U9B toggles at the end of 32 clock pulses (approximately 20 seconds) flip-flop U9A is reset, and U9B latches it in the off position. Transmitter keying is released, and all patch logic is reset to the standby condition. This sequence effectively prevents anyone from activating the patch and leaving a dial tone on the air for more than 20 seconds. The interval can be made shorter by changing the output of U5 which is connected to the clock input of flip-flop U9B.

If an acceptable telephone number is dialed within the time allowed the patch will remain activated until released by the *reset* digit or a logic 1 from the last stage of the timer. With the circuit shown this signal is generated after 512 clock pulses, or approximately 5 minutes. The 3-minute timer on the repeater is reset each time the caller keys his transmitter by contacts 5 and 6 of relay K2 so it will not time-out the repeater as long as he doesn't let the party on the other end of the telephone talk too long.

The degree and type of protection provided can be varied to suit the user and the requirements of the situation in which the autopatch system is to be used.

Security of access can be increased over that provided by the integral two-digit code by adding an external access module. This module can be used to control the *access clock* input to flip-flop U1, or provide a logic zero at the *external access control* terminal (this terminal should be grounded if no external access control is used). The external access module can also be wired to disable the transmitter while the two digits that activate the patch are received and decoded, helping to preserve their secrecy.

Another protective feature that can be varied to suit the user is the screening circuit. The simple circuit shown in fig. 6 will only insure that some number is dialed—in other words, that a dial tone is not left on the air until the repeater times out. As explained earlier, failure to dial at least one digit will deactivate the patch in 20 seconds.

Adding another AND gate and dual flip-flop as shown in **fig. 7** provides protection against long-distance calls. As in fig. 6, flip-flop U1A will be set by the first digit dialed after the access code. However, if this digit is a 1 (long distance) flip-flop U1B will also be set, and both U1A and U1B will latch. Flip-flop U2 will not be set, no *hold signal* will be generated, and the patch will deactivate itself as though no number had been dialed. Any digit except a 1 will set U1A without setting U1B; U2 will be set when U1A is reset by the next clock pulse and will generate the *hold signal.*

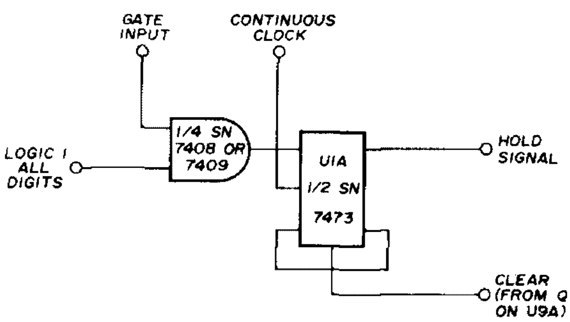


fig. 6. Simple screening circuit for the autopatch control logic. Dialing any number will hold the patch in the active condition (see text).

conclusion

Many other control variations are possible. A counter can be incorporated in the screening circuit to count the digits and deactivate the patch if too many or too few are dialed. External switching modules can be used to screen out the three- digit prefixes for weather reports, time signals and exchanges that are toll calls. The timer can also be connected to key the transmitter long enough for the dial tone to be heard, and then silence the repeater while the telephone number is being dialed. An external circuit can be added to override the five minute time limit and provision can be made to put the repeater in a standby mode at night—silent until the patch is activated by an emergency code. What you do with this circuit is truly limited only by your own imagination. I will welcome correspondence from readers interested in going beyond the capabilities of the circuit described here.

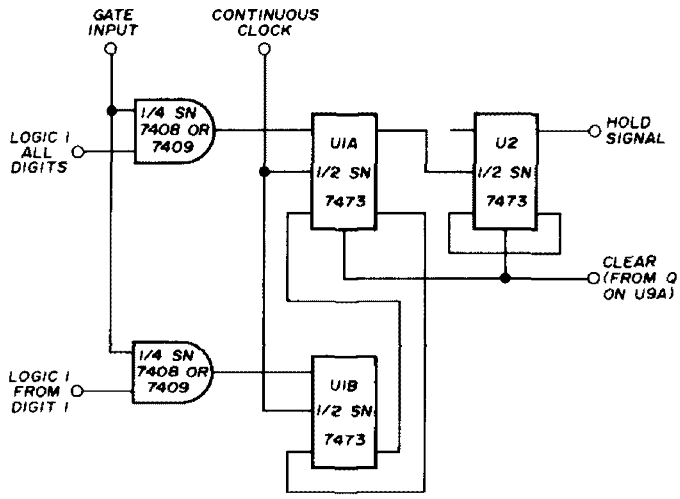


fig. 7. This screening circuit for the autopatch logic control system includes provision for locking out long-distance calls.

references

1. R.B. Shreve, W8GRG, "Superior Phone Patch," *ham radio*, July, 1971, page 20.
2. R.B. Shreve, W8GRG, "Integrated Circuit Sequential Switching for Touch-Tone Repeater Control," *ham radio*, June, 1972, page 22.

ham radio

5/8-wavelength two-meter antenna

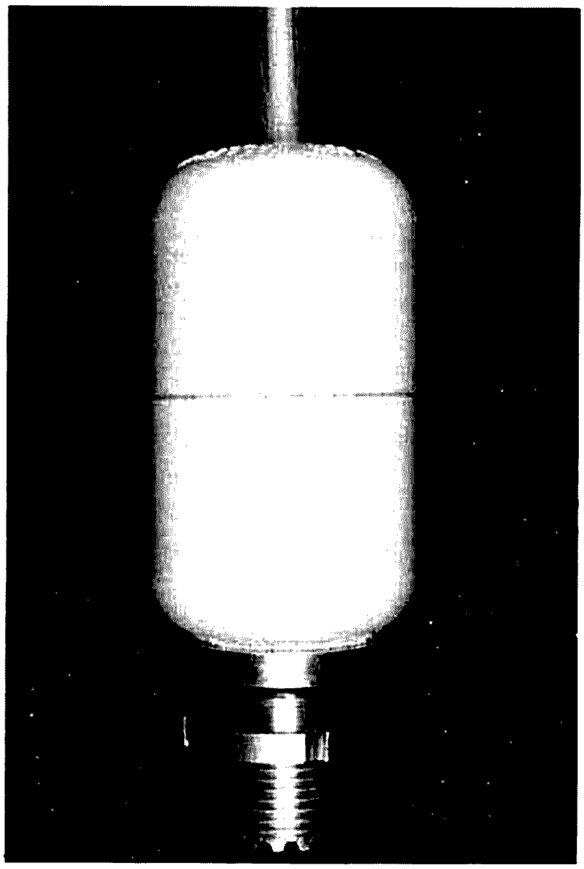
How to build
a low-cost
gain antenna
for your two-meter
mobile or base-station

Dave Sargent, K6KLO, 5343 Clovercrest Drive, San Jose, California 95118

The simplest way to match a 5/8-wavelength end-fed radiator to a 50-ohm feed system is to lengthen it by 1/8 wavelength with loading to make the antenna electrically 3/4-wavelength long.[1] The antenna will then present the same 50-ohm load as the familiar 1/4-wave whip. Radiation from the small loading coil will be almost nil, and the low angle radiation of a 5/8-wavelength radiator will be realized. Two-meter antennas of this type are available from several antenna manufacturers, but at rather dear prices. Described here is a 5/8-wavelength two-meter antenna which can be assembled for less than five dollars. The necessary materials are available from any hardware store, and standard hand tools are all that is needed for its construction. An electric drill is the only power tool required.

The 5/8-wavelength radiator is a replacement-type adjustable automobile antenna. The one I used was purchased from Allied Radio Shack for $1.39 (catalog number 12-1309). Many auto supply stores also carry these antennas, and any of them that will extend to 48 inches (1.22 meters) should be satisfactory.

fig. 1. Completed loading coil for 5/8-wave two-meter antenna. The type-uhf fitting at the base lends itself to a variety of mounting methods, and a replacement-type automobile broadcast antenna slips handily over the 5/16-inch brass rod on the top.



With the radiator portion of the antenna taken care of, all that is needed is a loading coil with a good mechanical base. Many approaches were considered, but none were going to be easy to construct. I finally tried the design shown in fig. 4, which went together successfully on the first try. All parts are labeled in the drawing. The 1/2-inch (12.7-mm) PVC plastic pipe caps cost about a quarter. Half-inch (12.7-mm) PVC pipe is usually available in ten-foot (3-meter) lengths for about fifty cents. Only two inches (5.1 cm) of pipe is needed, but the remainder may prove handy for some other project. Some number-16 copper bus wire for the coil, some epoxy, about two inches (3 cm) of 5/16 inch (7.9 mm) diameter brass rod, and a uhf coax connector (UG-266/U) or a piece of 3/8-24 (standard U.S. mobile mount) threaded brass stock round out the bill of materials. Although other coax connectors may be used the UG-266/U is best. A special PVC cement is available which is better than epoxy for gluing the plastic parts together.

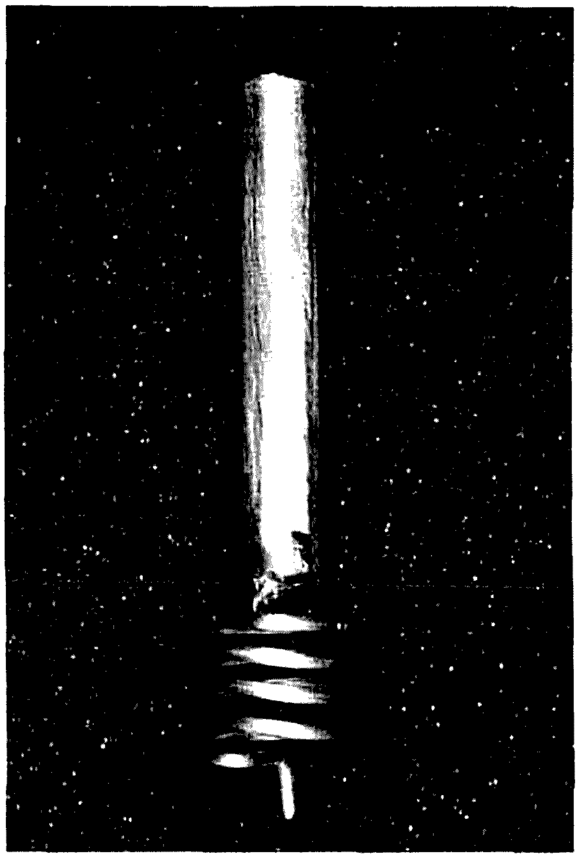
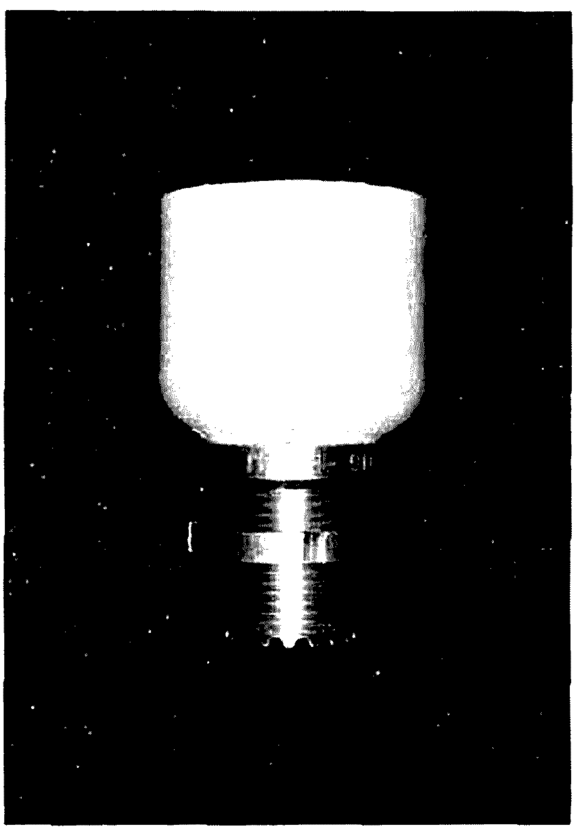


fig. 3. The loading coil is soldered to a hole drilled in the end of the brass rod, which then slips through a hole drilled in the top PVC cap.

fig. 2. PVC pipe cap with the UG-266/U coax connector installed.



coil assembly

To start the assembly, glue a two-inch (5.1-cm) piece of PVC pipe into one of the caps. The cap and pipe are then filled with about 3/4 inch (19 mm) of epoxy. This filling provides support for the 5/16-inch (7.9-mm) diameter rod used to mount the whip. With this assembly set aside for the epoxy to cure, a 9/16-inch (14.3-mm) diameter hole is drilled into the center of the other cap for the connector. A tapered hand reamer is satisfactory for making this hole if a large enough drill is not available. Screw the connector a short way into this hole, being careful to maintain alignment. Now heat the connector with a large iron or solder gun. When it is too hot to touch, grasp it with a pair of pliers and screw it into the cap. The heat will soften the PVC enough to allow this to be done and, after cooling, the connector will be molded into the PVC as shown in fig. 2.

After the epoxy has set, drill a 5/16-

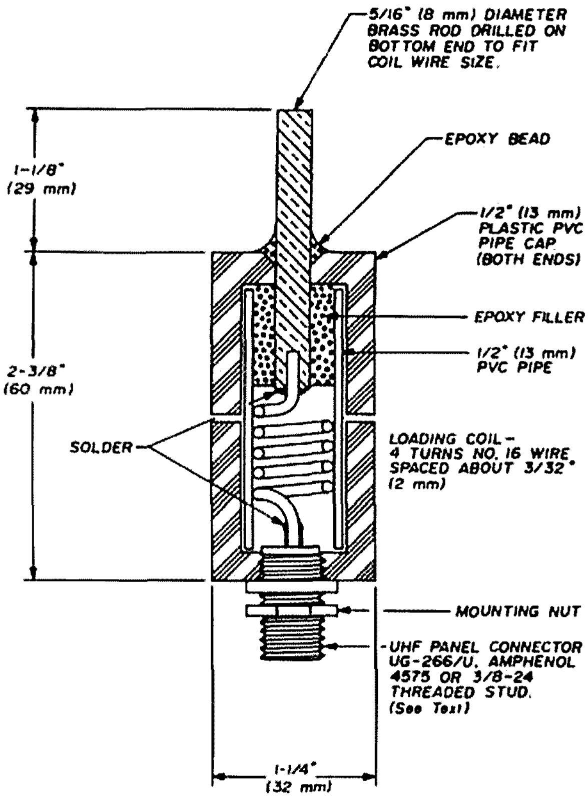


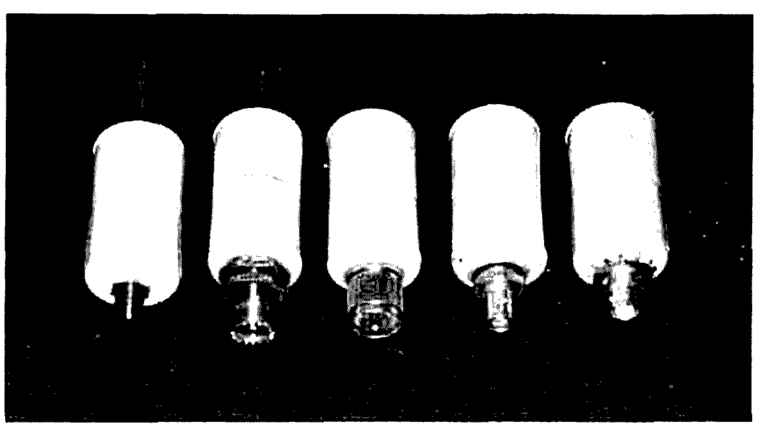
fig. 4. Cross-sectional drawing of completed loading coil assembly.

inch (7.9-mm) diameter hole in the center of the other cap. File this hole out slightly to allow easy insertion of the brass rod, bevel the top of the hole with a countersink or drill. Trim the PVC pipe to a length which will allow the caps to just come together when the halves are assembled. Wind four turns of number-16 wire on a form which will allow the finished coil to just fill the inside of the PVC pipe (about 9/16-inch (14.3-mm) diameter), and bend the ends of the coil to protrude radially from its center line.

Drill a small hole about 1/2-inch (12.7-mm) deep into one end of the 5/16-inch (7.9-mm) diameter brass rod and solder one end of the coil into it as shown in fig. 3. Solder the other end to the center terminal of the coax connector. Slip the top cap and PVC pipe assembly over the coil and rod to check the fit of the entire unit.

alternate mounting

In many mobile installations it is desirable to have the antenna screw into a standard mobile mount. In this case, the bottom cap should be drilled out with a size Q (about 21/64 inch or 8.4 mm) drill. A 3/8-24 threaded stud can then be heated and inserted in the same way that the connector was. A hole in the inside end of the stud should be drilled and tapped to provide a place for a solder lug for soldering the coil. Finally, a 3/8-24 nut should be run up the stud and tightened against the plastic cap. About 1/8 inch (3.2 mm) of the stud should protrude inside the cap, and some epoxy filler should be put around it to increase the strength of the base. This is necessary because the stud has 1/4 inch (6.4 mm) less diameter than the connector and might break out of the plastic if sufficient stress were placed on it.



Other connectors which have been used successfully are (left to right): 3/8-24 screw, UG-363/U UHF bulkhead feedthrough, UG-273/U UHF/BNC adapter, UG-492A/U BNC bulkhead feedthrough, and UG-911A/U panel jack (mounted from inside).

Before the assembly is sealed, its operation should be checked. For mobile installations, mount the antenna on an appropriate mobile mount. For a good match the base should be close to the car body, so do not use a large base spring.

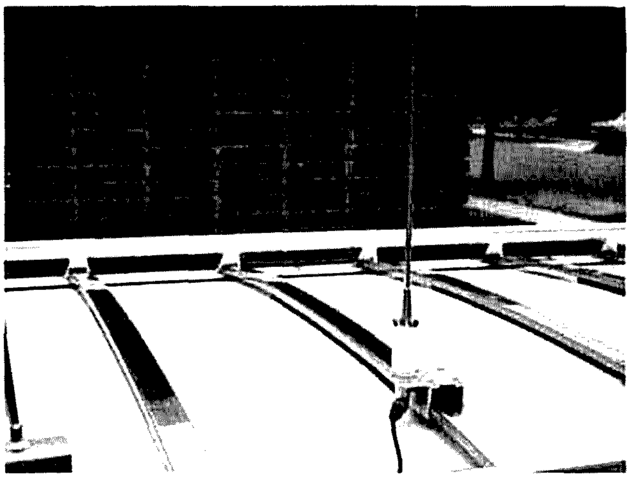


fig. 5. The 5/8-wave two-meter antenna installed on the roof of the author's station wagon. The set screws for mounting the whip have been replaced by thumb screws to permit quick removal for entering the garage.

For mast installations, the connector is mounted to an angle bracket through a hole, and four or more 20-inch (51-cm) radials must be added. In either case, adjust the whip to 48 inches (1.22 meters) and check the vswr. If it is not close to 1:1, adjust the whip length for minimum reflected power. If the whip needs to be lengthened the loading coil inductance needs to be increased; a shorter whip length means the coil requires less inductance. A whip adjusted to slightly shorter than 48 inches (1.22 meters) is acceptable, but whips of longer lengths should be avoided since undesirable high angle lobes will increase, and the low angle lobe will be weakened.

After any needed coil adjustments are made, coat the lower 3/4 inch (19 mm) of the brass rod with epoxy. The bottom cap may now be glued in place and a bead of epoxy placed in the beveled edge around the top cap as shown in the cross-section drawing, fig. **4**. The final assembly is now a rugged, air-tight unit equal in performance to expensive commercial gain antennas. It is well-suited to a variety of mountings such as the one used on the K6KLO station wagon in **fig. 5**.

reference

1. R.L. Crawshaw, "5/8 Wavelength Verticals," *73*, May, 1970, page 36.

ham radio

a vhf radio observatory

An interesting report on the relationship between the sun and the earth's weather, using data obtained at vhf

Ron Ham, Faraday, Greyfriars, Storrington, Sussex RH204HE, England

For many years prior to 1968 I had frequently heard solar radio noise whilst using vhf communications receivers for atmospheric studies. I realised these bursts of metre-wave radio noise were telling me that a solar event had taken place, and that a stream of complex particles may be heading toward earth. If the timing were right, these solar particles would enter earth's atmosphere and cause an aurora or disturb the natural state of the ionosphere, which would cause the normal propagation of a variety of radio signals to be upset.

With this in mind, I decided to build a radio telescope to give me prior knowledge of solar activity and to include this information in my propagation reports to the radio organizations.

the radio telescope

The aerial consists of four Yagis, each having four elements, mounted on a 10 x 6-foot (3 x 1.8-meters) wood frame, which is covered with a ½-inch (13-mm) wire mesh for the reflector. The whole reflector framework is hinged on its bottom rail so that its altitude can be adjusted periodically to keep the sun within its vertical beamwidth. The aerial faces south, and the earth drift principle is used for the azimuth adjustment.

The working frequency selected for the radio telescope is 135.95 MHz,

because at my location this frequency is free from terrestrial interference and the radio noise associated with sunspots can be very strong at this frequency.

The solar radio waves are fed to a crystal-controlled transistor converter mounted behind the aerial reflector. The intermediate frequency of 26 MHz, produced at the converter, is fed on underground coaxial cables to an i-f amplifier and detector. The dc voltage at the detector is amplified by a type 709 integrated circuit to drive a pen recorder.

The completed instrument was put into operation on June 1, 1968, and was soon producing results. Daily observations, which are controlled by a time switch, start when the sun enters the antenna beam at 1130 gmt and terminate at 1330 gmt; during this period, five feet (1.5 meters) of paper chart are produced from the recorder.

solar activity

As time passed, two features of solar activity emerged from the daily recordings. First was the burst of radio noise (fig. 1) and second was the noise storm (fig. 2). The individual solar burst may last only a few minutes, whilst the noise storm may continue for several days. The radio noise from a solar event is received 8.3 minutes after it has taken place, but the streams of nuclear waste, which are ejected by the sun at the same time, may take up to 40 hours before reaching the earth's orbital path.

The author is an amateur radio astronomer and a Fellow of the Royal Astronomical Society. For many years he has operated a radio observatory from his home in Sussex, England. Data from author Ham's observations have been supplied to the Radio Society of Great Britain, the International Amateur Radio Union (Region 1), and the British Astronomical Association. His equipment operates on 135.95 MHz, which favors the "active" sun noise. Although no technical data on the author's receiver are supplied with his article, it would appear that any good vhf converter could be used to duplicate the author's set for those who wish to expand their knowledge in this area. **Editor.**

The first "prize" observation came on November 1, 1968, when an Aurora Borealis was evident at the climax of a noise storm, which my telescope had been recording for several days. The great Auroras of March 8, 1970; August 5, 1972; and April 1, 1973, followed prolonged periods of solar activity, which had been recorded by my instrument. Throughout the five-year life of my radio telescope, many atmospheric disturbances

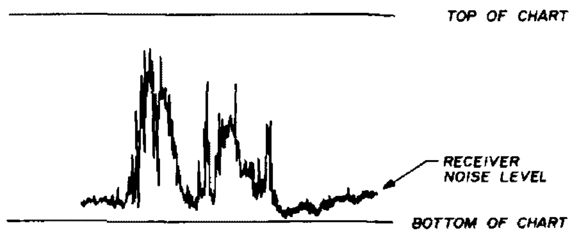


fig. 1. Four-minute burst of solar data recorded at 1308 gmt, June 29, 1973.

have been associated with solar activity. From the time I built this instrument, my friends in the radio world have been interested in its results and frequently have asked me for information to complement their studies.

objectives

The daily routine at my observatory consists of checking several vhf radio frequencies looking for the effects of Aurora, sporadic E, or tropospheric disturbances and then attempting to associate these effects with the results from the solar observations. In addition to the radio work, I record rainfall, humidity, temperature, wind speed and atmospheric pressure data for correlation with tropospheric openings. When possible I also log the "freak" weather disturbances that are large enough to justify the attention of the national news media.

solar activity and the earth's weather

Through keeping records of solar and atmospheric events I noticed that during or soon after I had recorded a noise storm

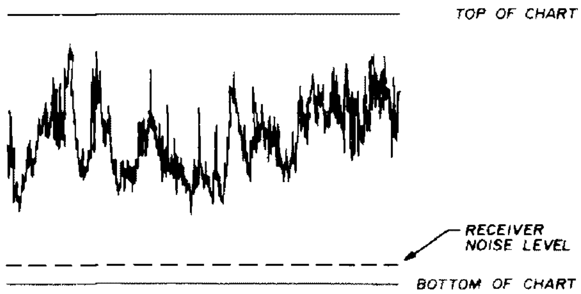


fig. 2. Eight-minute sample of solar activity recorded on August 7, 1972 during a storm that occurred August 1-9.

the news media were likely to report (somewhere on earth) a freak, or violent weather event, often with tragic loss of life and/or extensive damage to property. The following examples illustrate the typical weather news reports coinciding with recorded solar storms that set me thinking about a possible connection between these two natural events.

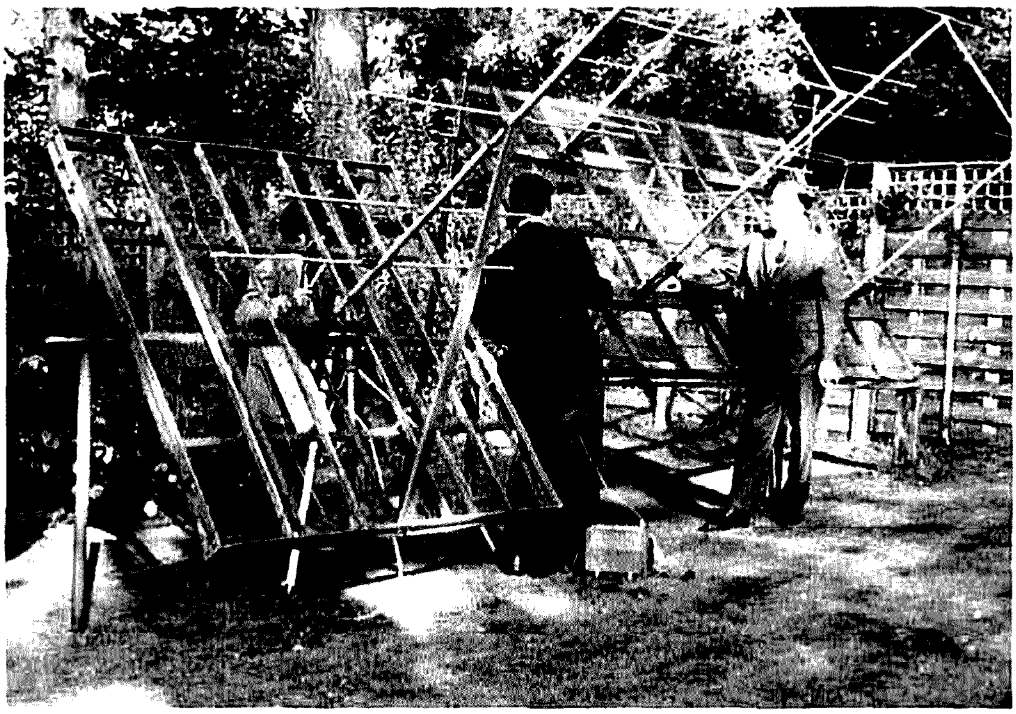
These are just a few of the reports I have gathered from newspaper items and radio news broadcasts; there may well be many more, or more detailed information in subsequent reports.

Scientific literature tells us that a connection between the active sun and the earth's weather has been known for many years, but the precise link has not yet been identified. Briefly, certain changes in climate, and in plant life, have already been associated with the eleven-year sunspot cycle. For approximately 400 years astronomers have systematically recorded the number of visible sunspots; and throughout these years, scientists have related many natural events to the existence of a large sunspot.

date of solar storm	date of news report
1970 - Nov. 11 to 22	**20th.** East Pakistan flood disaster.
1970 - Dec. 17 to 23	The unpredicted "White Christmas" in Europe with severe blizzards. (Some countries had snow for the first time.)
1971 - Jan. 28 to 31	**30th & 31st.** Floods in Poland and Mozambique. **Nine** inches (22.9 cm) of rain in one day was reported from Australia. Heavy rain and snow in parts of U.K. The Thames river was at risk of flooding because of severe gales in the North Sea.
1971 - April 9 to 18	**16th/17th.** Worst weather in 72 years experienced by Mount Everest climbers.
	13th. BBC news report that monsoons in E. Pakistan had started a month early.
1971 - July 13 to 19	**18th.** Freak rain storm in Seoul, Korea.
	20th. Hong Kong hit by worst typhoon for many years.
1971 - Aug. 18 to 27	**26th.** (Stop press). Storm havoc in Spain.
1972 - Feb. 12 to 23	**14th.** 100-mph (161-kmh) gales in southern France.
	18th. Freak storms in Australia.
	20th. Flooding in New York.
1972 - March 3 to 12	**13th.** Flooding in Peru.
1972 - June 15 to 22	**22nd.** Worst floods in American history; whole towns evacuated near New York; some parts declared disaster areas.
1972 - Aug. 1 to 9	**2nd.** Flooding in London.
1972 - Aug. 11 to 14	**11th.** Freak tornado reported in Holland.
1972 - Oct. 19 to 31	**24th.** Serious flooding in Australia.
	28th. Severe gales in Icelandic waters.
	20th. Severe flooding in Costa de Sol.
	31st. Gales in Somalia; many dead.
1973 - April 1 to 11	**2nd.** Hurricane—force winds in Holland.
	10th. Some parts of USA had snow for first time.
	14th. Storms in E. Pakistan; many dead.

solar radio waves

The association of radio waves with an active sunspot has been recognized only for about 40 years; prior to this, solar observations relied on what the eye could see. *Could it be that a simple amateur radio telescope has identified the particular sunspot activity responsible for stirring up the existing weather systems on earth?* After all, we know that solar particles, which are heralded by radio noise, can upset the ionosphere; so why, by some indirect means, can't they upset the troposphere?



Author's radio telescope aerials. At left is a BBC cameraman behind a 90-MHz aerial. A BBC interviewer and author (extreme right) stand in front of the 136-MHz aerial.

This sun/weather relationship phenomena caught the imagination of several friends, and it was suggested that my observations be placed on record. In August, 1971, an article containing some of these observations was published in the RSGB journal, *Radio Communication*. Since this article was published people have kindly sent me a variety of press cuttings about freak weather conditions for me to correlate with my solar recordings.

Amongst my friends, little more was talked about this matter until Jim Fisk raised the subject of geomagnetic effects on weather in his editorial of a recent issue of *ham radio*.[1] Like a shot from a gun an old friend, Brian Oddy, read this editorial and promptly contacted me to see if I had seen my copy of *ham radio*. Brian quickly pointed out that a solar storm could upset the geomagnetic field and in turn upset the weather.

Perhaps the editorial in *ham radio* has joined together two independant observations which, as Brian suggests, have provided another vital link in the chain of events between the sun and the earth's weather.

reference

1. Jim Fisk, W1DTY, "A Second Look", *ham radio*, April, 1973, page 4.

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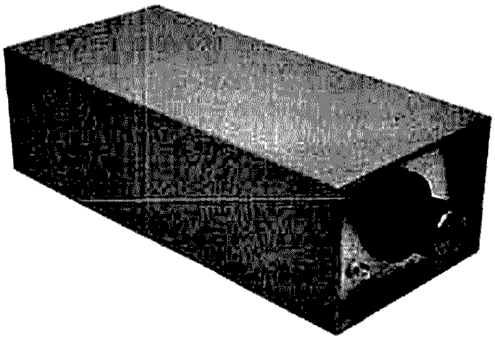
1. Joe Mikuckis, K3CHP, "Some Interesting Aspects of Increased Solar Activity," *ham radio*, June, 1968, page 21.
2. Hank Olson, W6GXN, "An Amateur Anemometer," *ham radio*, June, 1968, page 52. Short Circuit, *ham radio*, August, 1968, page 34.

ham radio



homebrew custom enclosures

A simple and economical technique for making handsome home projects

D.A. Contini, W4YUU, 4260 Middlebrook Lane, Orlando, Florida

How many times has one of your elegantly conceived projects turned into a "kluge" because the only housing you could find for it was an uninspiring standard Minibox? The satisfaction gained from building your own ham gear is determined by both performance and appearance, and it is the latter that suffers most severely from the limitations of the home workshop. As if Miniboxes weren't ugly enough, everything from office file-card boxes to a *Sucrets* can have also been used to house small electronic gadgets. Presented here, however, is a technique for fabricating your own attractive housing for that next project—one that requires only simple hand tools and is quite inexpensive.

materials

The key to this technique is rectangular cross-section extruded aluminum tubing. A shopping visit to an industrial aluminum supply house is your first step. You are looking for a stock of scrap pieces of 2 x 2-, 2 x 3-, 2 x 4- and 2 x 6-inch (approximately 5 x 5- through 5 x 15-cm) tubing. A wall thickness of 1/8

inch (3 mm) is best. Assorted lengths from 6 inches (15 cm) up are usually available from the scrap bin for a nominal price. At worst you should expect to pay $1.50 to $2.50 per foot (30 cm) if they have to cut it from stock. If they are very stuffy and will only talk in terms of 12-foot (3.6 meter) lengths you are at the wrong place. Besides the tubing, you will need some 0.032-inch (0.8-mm) to 0.062-inch (1.6-mm) sheet aluminum for the chassis and panels. Get a soft alloy like 6061-T4 so it won't crack when you bend it.

For finishing the metal, you will need a can of spray enamel. Sears and Roebuck epoxy enamel (catalog number 30F66258), made for painting appliances, works very nicely. Finally, you will need some 6-32 x 1/4-inch (6.35-mm) and 4-40 x 1/4-inch (6.35-mm) flat-head machine screws, and some 6-32 self-clinching nuts.*

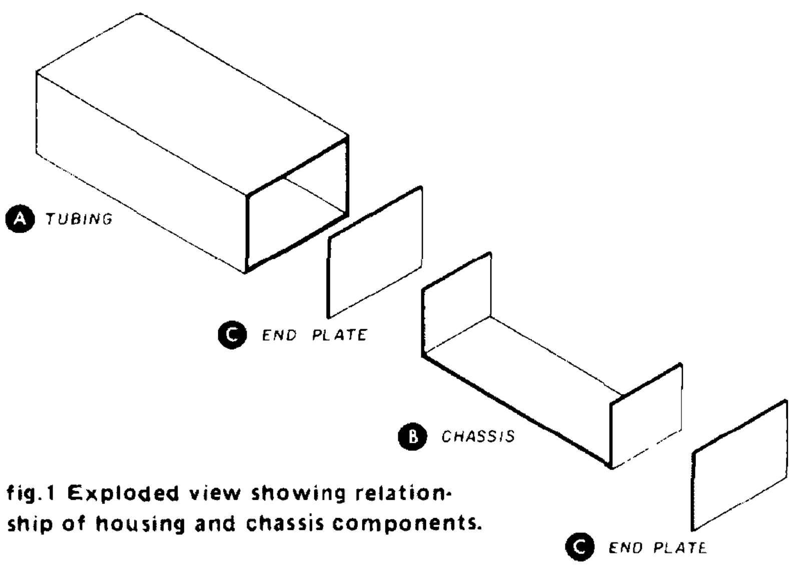


fig. 1 Exploded view showing relationship of housing and chassis components.

layout

The first step is to determine what length and size of tubing will be required for your project. A little "chess game" with the components will help you make this determination. As shown in **fig. 1**, a front and rear panel are cut from sheet aluminum to fit inside the tubing and a U-shaped chassis of the required length is formed from the same material. These parts are fastened together by the controls and connectors that are mounted on the front and rear panels. This assembly is then slid into a piece of aluminum tubing which is 1/2-inch (1.27-cm) longer than it. A 6-32 x 1/4-inch (6.35-mm) flat-head screw that goes through the bottom of the case and screws into a 6-32 clinch nut located at an appropriate point on the chassis holds the chassis assembly in place.

To recap, select a tubing size and length which is adequate to house the required components, making sure to allow room for all the controls and connectors to be mounted on the front and rear panels. Cut the front and rear panels to the inside dimensions of the selected tubing. Cut the chassis material 1/4 inch (6.4 mm) narrower than the inside width of the tubing to provide a 1/8-inch (3.2-mm) clearance inside the case. Form the chassis from a sheet which is as long as the required chassis plus twice the inside height of the tubing, minus 1/4 inch (6.4 mm). Cut the tubing 1/2 inch (1.3 cm) longer than the finished chassis and panel assembly. **Fig. 2** is a completely dimensioned layout for a 2 x 3 x 7-inch (5.1 x 7.6 x 17.8-cm) case.

fabrication

If you can arrange access to a sheet metal shear you can cut up a supply of panels and chassis stock for the sizes of tubing you have on hand. In the absence of a shear, cut the parts a little oversize with a hacksaw or saber saw and trim them to size with a file. Delay cutting the

*Available from Small Parts, Inc., 6901 N.E. Third Avenue, Miami, Florida 33138. Part number ON632-56, 10 for $1.15 or 25 for $2.50. $1.00 handling charge on orders under $5.00; postage is included. Catalog available.

tubing until the chassis and panels are assembled. Measure the panel height minus 1/8 inch (3.2 mm) from each end of the chassis material and bend these ends up 90 degrees. Clamp the front panel to one vertical end of the chassis so that the bottom edge of the panel is flush with the underside of the chassis, and drill the mounting holes for controls and terminals through both pieces at the same time. Repeat this process for the rear panel. Temporarily assembly the panels to the chassis using one or two of the controls to hold each end. Check to see that the assembled chassis slides smoothly into the tubing. Measure the length of the assembled chassis, and then cut the tubing 1/2 inch (1.3 cm) longer. Dress any rough edges with a file.

Select a point on the chassis where the clinch nut will not interfere with the intended component layout, and insert the chassis assembly into the finished length of tubing so that both panels are recessed equally.

Drill a 0.187-inch (4.8-mm) hole from the bottom of the tubing through both it and the chassis at this point. Remove the chassis assembly from the tubing and swage a clinch nut into the drilled hole in the chassis from the top. Countersink the matching hole in the tubing to accept a flat-head screw. Run a 6-32 x 1/4-inch (6.4-mm) flat-head screw through the bottom of the case into the clinch nut to lock the chassis into the tubing.

This completes the fabrication of the case. To mount circuit components to the chassis, use 4-40 x 1/4-inch (6.4-mm) flat-head screws inserted from the chassis underside, countersinking each hole to provide a smooth bottom.

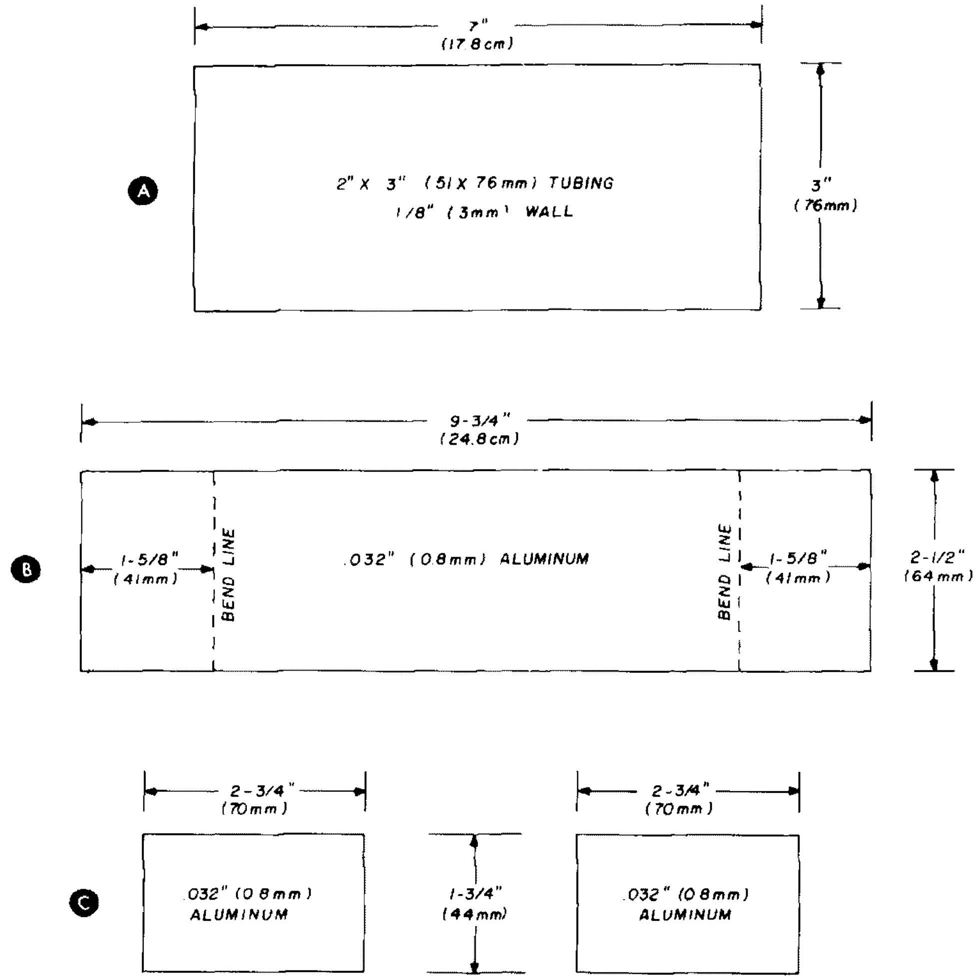
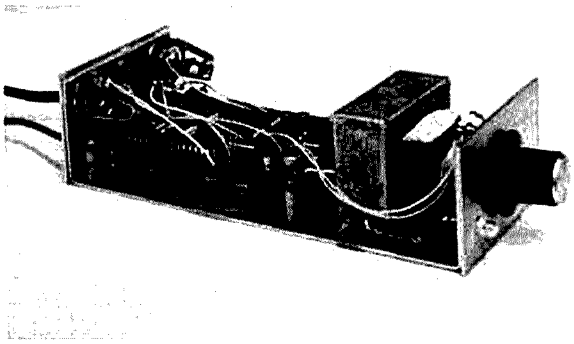


fig. 2. Dimensioned layout for components of an enclosure having a usable area of 2½x7x2 inches (6.4x17.8x5.1 cm).

finishing

Prepare the tubing for painting by giving it a light sanding. This removes surface scratches and provides a slightly roughened surface to improve paint adhesion. Next, wash the metal with hot water and detergent. To provide a completely



The Accu-Keyer removed from the case, showing how panel components hold the chassis assembly together and the technique for mounting the internal components.

grease-free surface do not touch the metal with your fingers. After rinsing with hot water, spray the case with two light coats of enamel, waiting about a half hour between coats. Carefully sanding the outside of the panels in a single direction will provide an attractive brushed-metal finish; alternatively, they too may be painted as the case was. The photographs show the popular WB4VVF Accu-Keyer packaged in this manner.

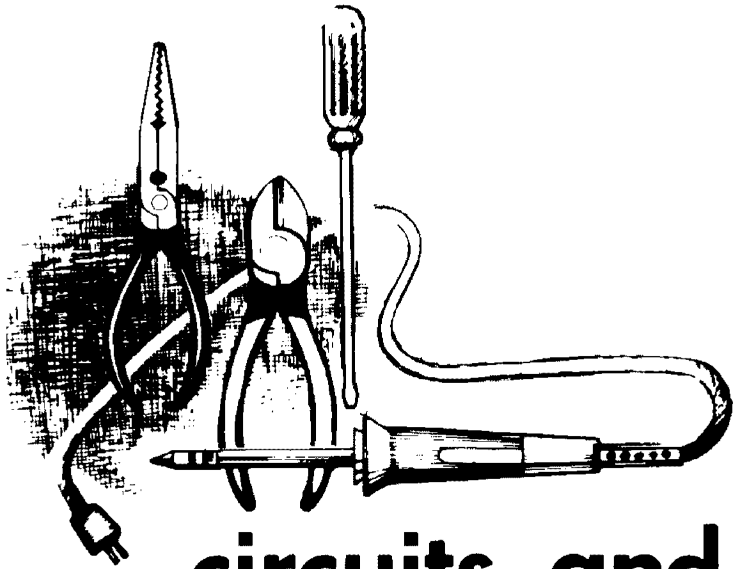
variations

As an alternative to the configuration described above, the tubing may be cut at an angle to give a "shadow" overhang at the top. The enclosure may be oriented with the short side of the tubing vertical, as in **fig. 1**, or with the short side horizontal when a tall narrow enclosure is preferred. Instead of enamel, you may wish to try a color anodizing if the facilities are available. Whichever way you go, you'll be pleased at the result.

reference

1. J.M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 91.

ham radio



circuits and techniques

ed noll, W3FQJ

solar energy

Solar power awaits a rebirth of man's wisdom. This form of energy cannot maim, endanger or exhaust the earth's treasures. The sun's energy is limitless and it will be around in quantity long after the earth has died. Nuclear and other forms of energy are plagued with one or more factors such as supply and demand limitations, unresolved technical problems, hazard, and long-term waste-disposal concerns. The use of solar energy presents no hazard to earth. There is no self-destruct potential. Is there any form of energy more free of pollution? Is there any form of energy or power more attractive and worth an all-out effort? Is not the sun the source of all energy on earth?

amateurs and solar power

How would you like to operate a solar-powered ham station? You can start at the QRPP level and work up—right up to the 1-kilowatt input limit. Today the initial installation is costly at high-power levels, but a solar power supply for low-powered equipment is reasonable and certainly will become more so as more and more amateurs go in this direction. Solar power activities are under way at W3FQJ and you will be kept informed.

some basics

First, let's put aside the notion that solar cells are only effective in bright sunlight and that only in the Southwest, with its endless clear days and bright sun, are solar-powered devices feasible. Actually, solar cells work quite efficiently even in considerable overcast. The secret is to match the installation with local average climatic conditions. The number of cells required for a specific application depends directly on the average weather conditions at the site. This simply tells us that in the East and Northeast, with their more generous portion of cloudy days, more solar cells are needed per given power demand than would be required in the sunny Southwest.

However, for each part of the country average solar energy levels have been measured for many years. Using this data you can select the proper number of cells required and add a few additional ones for good measure to accommodate a long sequence of cloudy winter days.

To make efficient use of solar light, the cell bed, **fig. 1**, must be tilted in a southerly direction. The average tilt angle corresponds with the latitude of the site (number of degrees that you are north of the equator). If your solar bed is mounted where it is readily accessible, you can make minor adjustments in this

angle in spring and fall to compensate for the somewhat different path taken by the sun across the heavens between dawn and dusk. (More exactly, in an astronomical sense, compensation is made for the changing tilt angle of the rotating earth, the sun being in a fixed position.)

A solar cell system or, more precisely, a solar energy or light energy converter, can be designed to *deliver a certain* amount of power in daylight. However, any solar energy system, to be worthy, should also be capable of taking care of

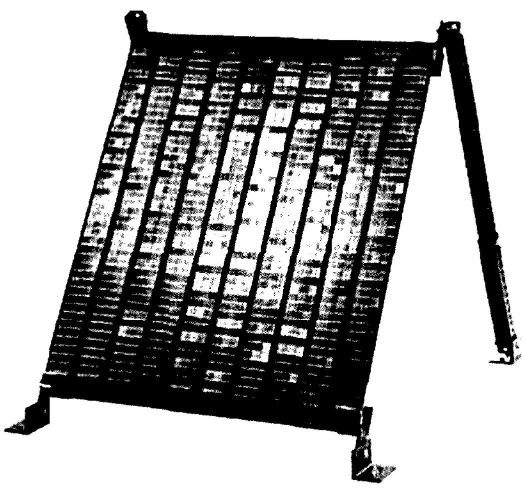


fig. 1. Light-energy converter which could be used for a solar-powered amateur radio station. (Photo courtesy Spectrolab.)

nighttime needs. In terms of electrical power energy this can be handled by chargeable batteries. Thus a 24-hour, all-year system requires a light energy converter and a battery pack. Such a system can be combined into a completely self-sustaining and unattended installation.

An alternative manual system would be one that would permit daytime operation from the solar converter; nighttime operation would be by battery. However, the battery's capacity would have to be such that it could handle the nighttime operating hours and power demand. This battery would be charged to desired capacity sometime during the day using the light energy converter.

The two big factors in planning a successful solar power system involve the power requirements and how these demands can be met by a solar converter. The criteria for the latter were mentioned previously. Next consider how the average power demand can be estimated. A voltage requirement must be met. This is accommodated by the number of individual solar cells that are connected in series for the solar converter. Load current must be averaged on the basis of day-night operations and then over a long-term solar-year basis.

A practical approach is to determine the ampere-hours required per day. This figure can be averaged for the solar year. Current demand evaluation should also be established for nighttime operation as an aid in choosing a battery source with the required ampere-hour capability. Peak power is also a consideration in determining the peak current demand at any time to make certain the solar converter and battery pack are capable of delivering this current level. After the power requirement evaluations are made, the solar energy converter is designed to deliver at least this amount of voltage, current and power based on the solar conditions at the mounting site.

A small safety factor may be advisable although initial cost can be kept down by not exaggerating possible power requirements. It is no great problem to add additional solar beds later if an increase in power demand is anticipated.

A functional diagram of a complete, unattended system is shown in fig. 2. Only a few basic components are needed in addition to the light energy converter and batteries. A charging diode is needed to present a conducting path between the converter and the batteries to be charged. This diode acts as a one-way path, preventing the batteries from discharging into the converter when they have been fully charged. This completes the power source when using lead-acid storage

batteries. An additional voltage regulator may be advisable to prevent overcharge of nickel-cadmium batteries. Conductors, of course, should have adequate size to minimize IR losses. Too great an IR drop between converter and battery results in inadequate charging current.

like batteries, to obtain a desired converter voltage. Likewise, the current capability is increased by connecting the cells in parallel. This series-parallel grouping of cells into an array permits the construction of a bed of solar cells that will supply the voltage and current needs

table 1. Standard 12-volt LEC™ solar-power arrays manufactured by Spectrolab.

array model number	approximate dimensions (inches)	current output	amp-hours generated under various per day solar conditions[2] poor	average	good	approximate weight (lb)
12V 300 mA	37 x 3 x 3	300 mA	1.1	1.3	1.6	3.8
12V 600 mA	37 x 6 x 3	600 mA	2.2	2.6	3.1	9.3
12V 900 mA	37 x 9 x 3	900 mA	3.3	3.9	4.7	14.0
12V 1.2 A	37 x 12 x 3	1.2 A	4.4	5.2	6.3	18.6
12V 1.5 A	37 x 15 x 3	1.5 A	5.5	6.5	7.9	23.3
12V 1.8 A	37 x 18 x 3	1.8 A	6.6	7.8	9.4	28.0
12V 2.1 A	37 x 21 x 3	2.1 A	7.7	9.0	11.0	32.6
12V 2.4 A	37 x 24 x 3	2.4 A	8.9	10.4	12.6	37.3
12V 2.7 A	37 x 27 x 3	2.7 A	10.0	11.7	14.1	41.9
12V 3.0 A	37 x 30 x 3	3.0 A	11.1	13.0	15.7	46.6
12V 3.6 A	37 x 37 x 3	3.6 A	13.3	15.6	18.9	55.9
12V 4.2 A	37 x 43 x 3	4.2 A	15.5	18.2	22.0	65.2
12V 4.8 A	37 x 49 x 3	4.8 A	17.7	20.8	25.2	74.5
12V 5.4 A	37 x 55 x 3	5.4 A	20.0	23.4	28.3	83.8
12V 6.0 A	37 x 61 x 3	6.0 A	22.2	26.0	31.5	93.1
12V 6.6 A	37 x 67 x 3	6.6 A	24.4	28.6	34.7	102.4
12V 7.2 A	37 x 73 x 3	7.2 A	26.6	31.2	37.8	111.7

1. Minimum current output under Standard Test Conditions (STC) Intensity = 100 mw/cm^2; Temperature = 0°C to + 60°C.
2. Usable energy generated for use in a solar power supply system with lead-acid storage batteries under typical conditions and based on annual mean solar radiation data for various locations in the contiguous United States.

Silicon solar cells are light-sensitive semiconductor devices. P- and N-type impurities are added to a basic silicon crystal. For example, the basic wafer can be a P-type semiconductor. An N-type layer can then be diffused a certain depth into the wafer, **fig. 3**. A P-N junction is formed between the two layers. When light is directed onto the junction, electron and hole carriers are formed by the impacting light photons. The hole carriers move to the N-type region; electron carriers, to the P-type region. This motion of charges across the junction constitutes an electric current. The path is completed through the external circuit.

A single solar cell has only a small voltage drop and a limited current capability. Cells are connected in series, just

of the light-energy converter. All of these must then be assembled in a durable frame and support structure, **fig. 1**, including output terminals.

The assembly must be made as impervious as possible to weather and other environmental extremes. A bracket arrangement is needed for obtaining the proper tilt of the array at the mounting site. The solar cells themselves are not exposed to the elements because they are protected by an efficient transparent coating that provides proper diffusing and channeling of the arriving light energy. Even with considerable icing the conversion efficiency remains high.

The directional diode is usually a silicon type, selected with proper voltage rating and adequate current-carrying

capability. Keep the diode voltage drop as low as possible.

The chart of **table 1** shows the standard 12-volt light energy converter array made by Spectrolab.* The first unit, about three-feet (92-cm) long and three-inches (75-mm) wide, supplies 12-volts at

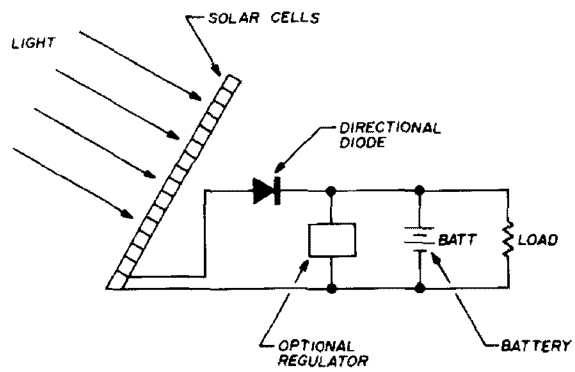


fig. 2. Circuit for a basic solar energy converter and battery system. Battery is recharged during the day, provides power at night.

300 milliamperes. This is the minimum current under standard test conditions. This standard test condition corresponds to the solar intensity at noon on a clear day when the temperature is 77°F (25°C). In designing the system it is necessary to derate current values on the basis of higher operating temperatures. This is done in the design of the particular array module to be used at a given site. The three-column, ampere-hours specifications are interesting since they represent the useful ampere-hours that can be supplied by a solar power supply system using lead-acid storage batteries. These typical conditions are based on annual mean solar radiation data for various locations in the continental United States.

batteries

Battery quality is an important consideration. Low-cost lead-acid cells can be charged by solar converters of adequate size. However, for all-day, all-year, uninterrupted service, batteries should be selected more carefully if initial cost and efficient operating conditions are to be achieved. Inexpensive types have a high cell-discharge rate and perhaps a two-to-four year potential life. However, high-quality lead-acid storage batteries are made by various manufacturers. Some of these have a self-discharge rate as low as 10 to 15 percent per year, and have a lasting capability of 10 to 15 years in a properly designed solar power supply. Nickel-cadmium batteries do very well.

The storage capacity of the battery should be based on peak daily use, considering also the number of days such a system may need to operate at reduced solar intensity. In an optimum system it is customary to incorporate approximately seven days of reserve battery capacity so as to preclude system failure under several days of very low light levels.

Some quality lead-acid batteries have a charging efficiency of 95%. This means that 95% of the power *delivered* to the battery ends up as charge. Thus, considerably less power must be delivered from the solar energy converter for a given level of battery charge.

amateur requirements

Radio amateur applications in general would not be nearly so stringent. There is

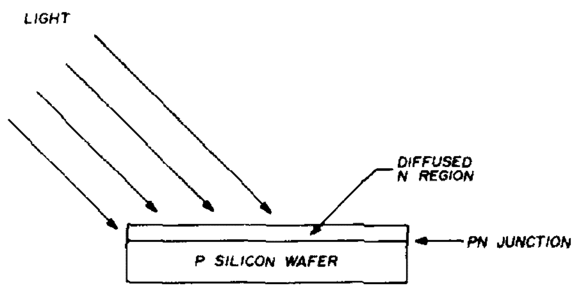


fig. 3. Construction of the basic silicon solar cell.

no need for all-day, all-night and all-year uninterrupted capability. Perhaps the only exception would be for the operation of a solar-powered fm repeater. Consider how many hours per day you operate your station. What is the longest span of continuous operation, on the

*Spectrolab, 12500 Gladstone Avenue, Sylmar, California 91342.

average? Do you operate every day? When you are on the air what is the ratio of your transmit time (when power is drawn) compared to your receive and listening time? All of the above factors mean that the power requirements of any solar energy converter for the usual amateur radio application can be much more modest than the restrictions of commercial use.

Is the average amateur on the air more than 2 to 3 hours per day? Even though a station may be on the air each day, the total hours may actually be less than 12 per week. No doubt the actual transmit time is less than half of the total amount. It would be interesting to go through your log and determine just how many operating hours you have per month. When your total operating hours are this modest, solar power could be quite feasible, even for a 200-watt PEP sideband transceiver.

Some sample figures will help to clarify power demand. Assume a solid-state transmitter that would draw 8 amperes from a 12-volt source. If the transmitter were on continuously drawing maximum power, a 60-ampere/hour storage battery would operate the unit for 5-6 hours on one charge. However, the actual transmit time is perhaps less than 50% of total operating time. Hence the battery would not discharge completely in six hours (complete discharge is to be avoided). Furthermore, using sideband transmission, the heavy demand is only made on those occasional modulation peaks. It is apparent that a good number of operating hours are feasible with a single charge of a good quality 12-volt battery.

If the battery has a 120-ampere-hour capability, a single charge might be adequate for a good number of operating hours per week. A trickle charge from a solar converter could readily maintain battery charge. In fact, you would probably not even require continuous connection to the solar converter. Charge time during two or three clear days of the week would be adequate in most cases.

Admittedly, at present production the cost of such a solar converter would be substantially higher than a conventional mains-powered battery charger. However, with proper care the life of such a converter would be 15 years or more for the present state of the art. It offers a practical means for conserving precious electrical energy and would be especially useful for those locations where no electrical power is available for a battery charger. Here is an answer for those amateurs in countries plagued by electrical black-outs and brown-outs.

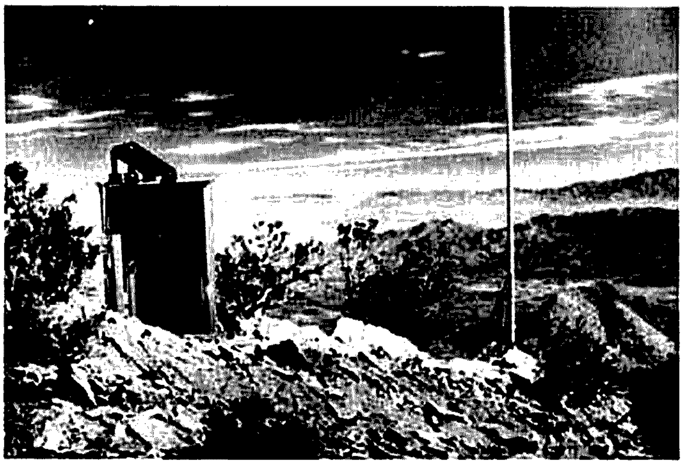


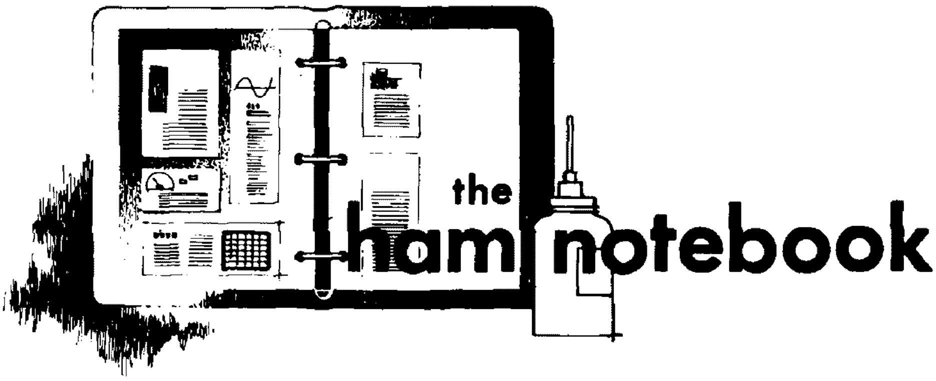
fig. 4. Solar power site installed by the Bureau of Land Management at Clark Mountain, California. (Photo courtesy Spectrolab.)

low-power transmitters

At the QRP and QRPP levels it costs very little to get involved with solar energy conversion and gain a knowledge of the technique. Off-the-shelf photocells and associated components can get you started at the 100-milliwatt level. Daylight operating power is no problem at all. The addition of a nickel-cadmium battery (very low-powered in terms of ampere/hour charge and therefore inexpensive) will give you nighttime operation as well.

Going up to 1 or 2 watts involves very little additional cost. An upgrade to 5 to 10 watts requires a few bucks but adds fun, satisfaction and versatility, even to powering a low-powered sideband transceiver.

ham radio



precision voltage supply for phase-locked terminal unit

A previous article in *ham radio* described a phase-locked loop RTTY tuning unit which requires a stable, well regulated source of +12 volts.[1] Author W4FQM suggested using a commercially available power supply and described it as, "quite a buy for only $38.00." Well, hold on to your money fellows, and read on.

Fig. 1 shows the schematic of the precision supply. A full-wave bridge rectifier feeds the regulator. The value of C1 provides sufficient filtering with the regulator rejecting any remaining ripple. Resistors R1 and R2 determine the output voltage based on the following manufacture data sheet equations:

$$R1 \approx (2V_{out} - 7) \text{ kilohms}$$

$$R2 = 6.8k$$

Resistor R_{sc} and transistor Q1 provide short-circuit protection for the regulator. When the output short-circuit current

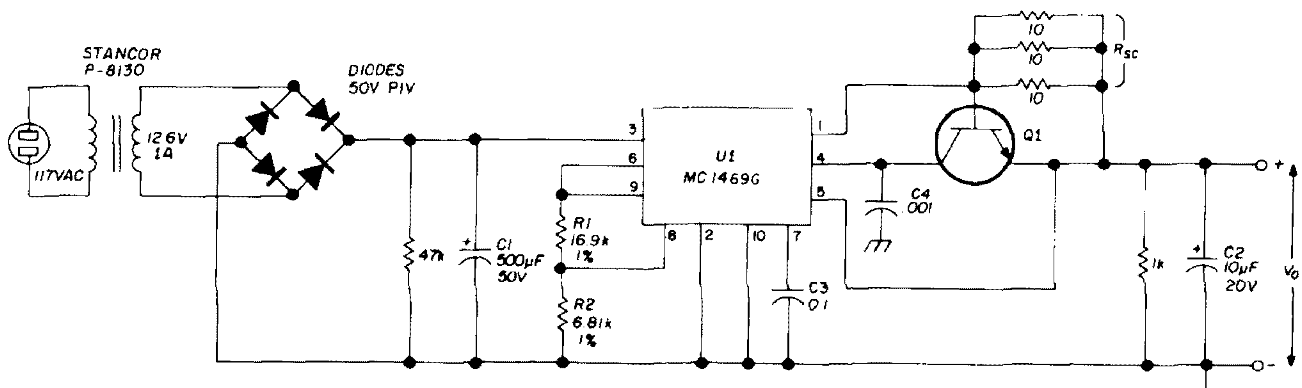


fig. 1. Precision 12-volt regulated supply. U1 may be a Motorola MC1469G or HEP C6049G.

The power supply described here uses an IC voltage regulator, Motorola MC1469G, and two precision 1% resistors as special parts. Everything else should be available in the average RTTY enthusiast's junk box. Except for these specific parts, other values and part types are not critical.

(I_{sc}) creates a voltage drop across R_{sc} large enough to turn Q1 on, the regulator output is limited by the saturated collector-emitter across pins 4 and 5. The value of R_{sc} is determined by the equation $R_{sc} \cong (0.6/I_{sc})$ ohms and $C_2 \leqslant (250/R_{sc})\ \mu F$, where I_{sc} is expressed in amperes and C2 is 250 μF maximum.

performance

The design I tested uses the values shown in the schematic. With the three 10-ohm resistors in parallel, the value of I_{sc} measured was 200 mA. For a nominal load current of 50 mA the output voltage was 12.235 volts. Changing this load current ±10 mA caused the output to

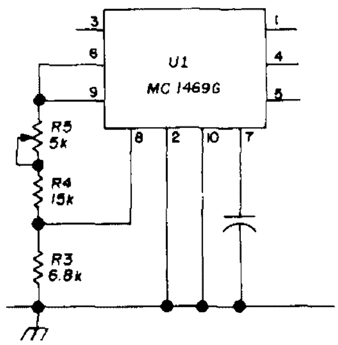


fig. 2. Alternate resistor configuration for the 12-volt regulated supply. R5 is a miniature multiturn potentiometer.

vary ±1 millivolt. This corresponds to a load regulation of ±0.008%. The power supply showed a 0.01% change in output voltage with a 7% change in the input ac voltage. Both of these characteristics more than satisfy the ±0.1% regulation requirement of the PLL.

The output voltage was designed for a nominal 12.0 volts and measured at 12.23 volts. This is explained by the approximately equals sign in the equation for resistor R1. The absolute value of the output is not critical so long as the output is stable.

construction

Layout and construction of the circuit is not critical except that the manufacturer *recommends the .001-µF capacitor* on pin 4 of the MC1469G IC must have short lead lengths for regulator stability. Vector boards and point to point wiring is a lot easier than trying to design a one-time printed-circuit board. Sockets were used for the IC and Q1 but are not necessary. Just make sure when soldering the leads that a heatsink is used and all soldering is done quickly to avoid overheating.

The alternate configuration shown in **fig.** 2 can be used in place of the 1% metal-film resistors for R1 and R2. Resistors R3 and R4 are carbon composition resistors and R5 is a multiturn trimpot or a fixed composition resistor. Specific values of R4 and R5 are not important so long as they can be varied over the range of desired output voltage. The use of carbon composition resistors will degrade the long term stability of the supply but should not significantly effect TU performance.

Transistor Q1 is any general purpose npn silicon transistor. Rectifier diodes should be greater than 50 volts PIV. A clip-on TO-5 heatsink is used as a precaution because the regulator dissipates approximately 300 milliwatts.

Elliott Lawrence, WA6TLA

reference

1. Ed Webb, W4FQM, "Phase-Locked RTTY Terminal Unit," *ham radio,* January, 1972, page 8.

Collins S-line power supply mod

I found that I had to readjust the idling current potentiometer in my Collins 516F-2 S-line power supply quite frequency due to changes in the 117-volt ac power line. Connecting two 36-volt 400-mW, 5% zener diodes (1N974Bs) in series from the junction of R8 and R9 to ground as shown in **fig. 3** takes care of the problem nicely.

Ralph Cabanillas, Jr., W6IL/CT1HO

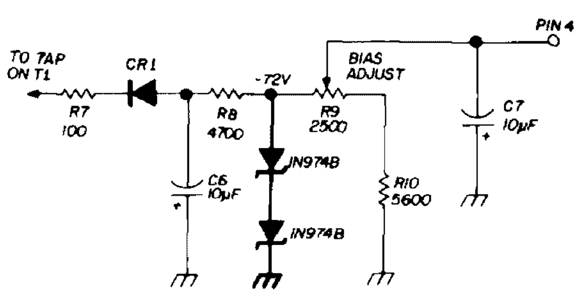
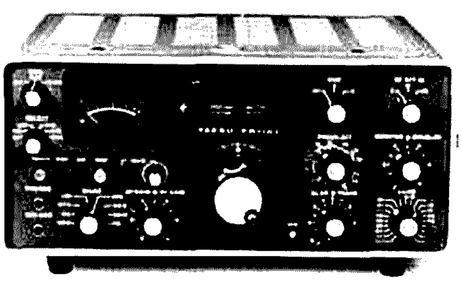


fig. 3. Negative bias portion of Collins 516F-2 power supply, showing two 36-volt zener diodes added to regulate the bias.

new products

communications receiver



Yaesu Musen has just introduced a new solid-state communications receiver, the FR-101S, with provision for all-mode reception on twenty-one 500-kHz amateur and shortwave bands from 160 through 2 meters. This new receiver is designed to be used in transceive, if desired, with the new FL-101 transmitter which is to be introduced in the near future. New solid-state technology, with features such as a doubly-balanced mixer, offer excellent rejection of cross-modulation and intermodulation interference. The FR-101S, which copies a-m, fm, ssb, CW and RTTY, has less than 100 Hz drift in any 30-minute period after warmup. Sensitivity is 0.5 μV for 10-dB signal-to-noise ratio on ssb and CW, 1.0-μV for 10-dB signal-to-noise ratio on a-m and 12 dB SINAD on fm, comes complete with 2.4-kHz, 4.0-kHz and 0.6-kHz crystal filters (1.5-kHz, 12-kHz and 45-kHz filters are available as optional accessories). Image rejection is 60-dB minimum and audio output is two watts into 4 ohms.

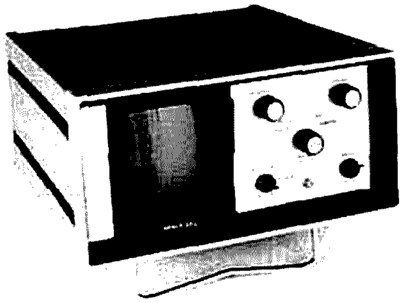
The FR-101S, which tunes the 160-, 80-, 40-, 20-, 15- and parts of the 10-meter band in standard trim, can be used on six and two meters and other shortwave bands with optional accessories. Priced at $499, this new receiver is available from Yaesu Musen USA Inc., 7625 East Rosecrans Avenue, Unit 29, Paramount, California 09723. For more information use *check-off* on page 94.

scanner-monitor servicing data

A new edition of Howard Sams' popular *Scanner Monitor Servicing Data* is now available. This new volume, third in a series, covers many of the popular Regency units including the MT-15S, TME-16H/L, TME-16H/LH/U, TME-16H/LL/U, TME-16H/LM/U, TMR-1H, TMR-1L, TMR-4H, TMR-4L, TMR-8H, TMR-8L, TMR-8H/LH, TMR-8H/LL and TMR-8H/LM. Included are schematics, parts lists and complete service adjustments. Other scanners covered in this new volume are the Electra Jolly Roger; Johnson Hi/Lo Duo Scan, UHF/VHF Duo-Scan, UHF Mono-Scan and VHF Mono-Scan; Midland 13-914; Pearce-Simpson Cherokee 8+8, Cheyenne 8 (PR-78) and Comanche 16 (PR-160). Available for $5.95 postpaid from Ham Radio Books, Greenville, New Hampshire 03048; order book number SD-3. Earlier volumes SD-1 and SD-2 ($4.95 each), which provide the same sort of complete servicing data on other scanner units, are also available.

More Details? CHECK—OFF Page 94

slow-scan monitor



Venus Scientific has introduced their new SS2 Slo-Scan Monitor. This Monitor is the second generation of Slow Scan with many features not previously available on the market. These features include Accu-Sync,TM a diagnostic and tuning aid which converts the SS2 Monitor to an oscilloscope by the flip of a switch, LED sweep indicators for ease of servicing, camera adapter provision which enables you to take Polaroid photographs right off the air with the P-1 Camera Adapter and simplified independent controls. trols.

The SS2 Monitor's picture size is 4-7/16-inch (11.3-cm) diagonal, 3–1/4 x 3-1/8 inch (8.3 x 7.9 cm) with 128 lines. It has a 15-Hz line rate and a 8-1/2-second frame rate. Video input modulation is fm, 1200 to 2300 Hz. Complete details may be obtained from Venus *Scientific, Inc.*, 399 Smith Street, Farmingdale, New York 11735, or use *check-off* on page 94.

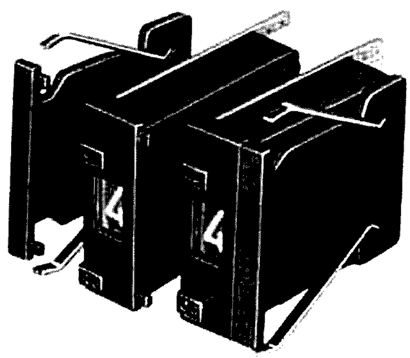
rf clipper

Holdings of England has introduced a unique new rf clipper for use with the Yaesu FT-101, Mark 2, that is used on both transmit *and* receive. The extra sideband filter provides a noticeable

improvement in adjacent channel selectivity on receive, and since the gain is added after the filter, this gives improved agc action with an *apparent* improvement of cross-modulation performance. The no compromise, all fet design uses a high quality input filter that matches the ssb filter supplied with the FT-101, Mk 2. A diode switch on the output control, in conjunction with a controlled fet stage, adjusts the gain independently on both receive and transmit. On receive the gain is set to give a boost of approximately two S-units.

This rf clipper, which should not be confused with normal clippers which are often no better than a good quality microphone, is wired to an octal plug which fits into the vfo socket. *No modification* of any kind is required to the FT-101, Mk 2 (device cannot be used on earlier FT-101). Normal operation without the clipper is restored in seconds by inserting a shorted octal plug. Units will be shipped from England via airmail. For more information write to Holdings Ltd., 39/41 Mincing Lane, Blackburn BB2 2AF, England, or use *check-off* on page 94.

bi-directional code switches



Alco Electronics' new SMC Series bi-directional pushbutton code switches maximize reliable performance in a small package. These modules occupy a panel area only 0.945-inch high by 0.3-inch wide, yet the position indicator numerals are an easy-to-read 0.2-inch high. The compact size is ideal for compact portable and mobile applications. Push-

buttons marked + and - allow the operator to advance or reverse numerical sequencing.

Available standard codes include conventional BCD (8-4-2-1), BCD with compliment, and decimal (1-of-10) formats, all 10-position types. The numerals of the visual readout are 0 through 9, corresponding to the electrical output codes. Electrical contact surfaces are gold plated for long, trouble-free life.

Possible applications for SMC switches include control of frequency synthesized tuners, channel selectors, and preset counters and timers. For special applications, such as 2-meter fm tuners, dummy switches (nonfunctioning, but identical in appearance) are available, with fixed numeral (e.g., "1" or "4"). Switches with limit stops, restricting the range of operation, are also available. For further information use *check-off* on page 94, or write to ALCO Electronic Products, Inc., 1551 Osgood Street, North Andover, Massachusetts 01848.

world radio and tv handbook

When a specialized handbook like this has gone into its 28th edition, there's very little that's "new and exciting" that can be said. The *World Radio & TV Handbook* (popularly called the *WRH*) is the only complete and comprehensive directory on radio broadcasting throughout the entire world—from Afars to Zambia. Updated during the latter part of 1973 for use during 1974, the *WRH* tells it all: stations, callsigns, frequencies, schedules, languages, power, etc. If you want the shortwave schedule for a certain country, the *WRH* is the place to find it. If you tune in a new station and wonder which one it might be, the List of All Shortwave Broadcasting Stations will give the information. No casual—and certainly no serious—shortwave listener is ever without a copy of the latest edition of the *WRH*. 408 pages, softbound, $7.50 from Ham Radio Books, Greenville, New Hampshire 03048.

impedance bridge

The Amateur Products Group of Delavan Electronics, Inc., has announced the development of Dela-Bridge I, which is designed to analyze antenna characteristics and simplify adjustments. This new instrument, when excited by a grid dip meter or low-power transmitter, quickly analyzes existing antenna and feedline characteristics, tuning and loading coils, and filter and interstage coupling networks. A direct readout allows easy adjustment for optimum performance.

Frequency range of the Dela-Bridge I is 50 kHz to 250 MHz with a resistance range of zero to 500 ohms, balanced or unbalanced, logarithmic scale. Excitation requirements are one mW to two watts maximum. An internal nine-volt battery provides power to the instrument, which has an accuracy of ±3% at 50 ohms. The readout, which is not frequency sensitive, provides complete null and reactance determination and the internal integrated circuit amplifier allows use with low-signal inputs.

Guaranteed by Delavan Electronics for one year, the Dela-Bridge I is available for $39.95 plus $2.50 for air mail and handling costs. A ready-to-assemble kit is available for $29.95. For more information write to Delavan Electronics, Inc., 14441 North 73rd Street, Scottsdale, Arizona 85260, or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

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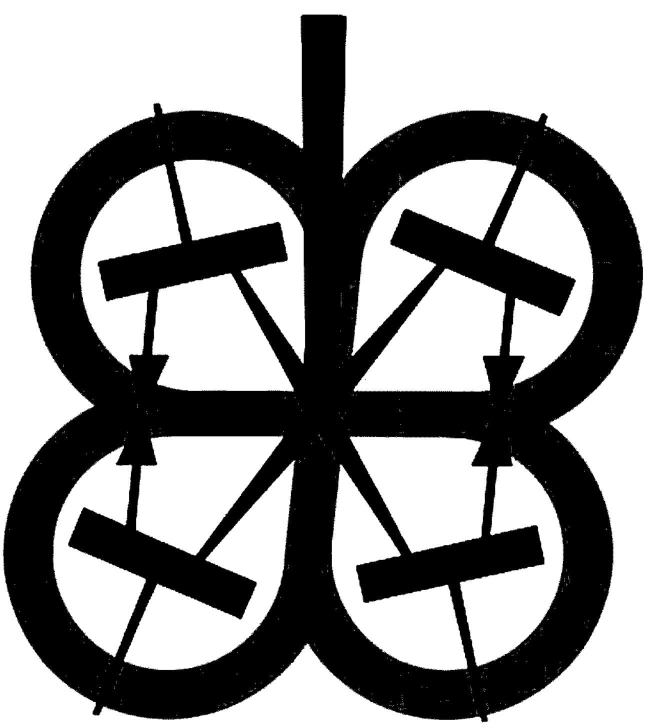
ham radio magazine

AUGUST 1974

high-power solid-state LINEAR POWER AMPLIFIER



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ham radio magazine

contents

a second look

by jim fisk

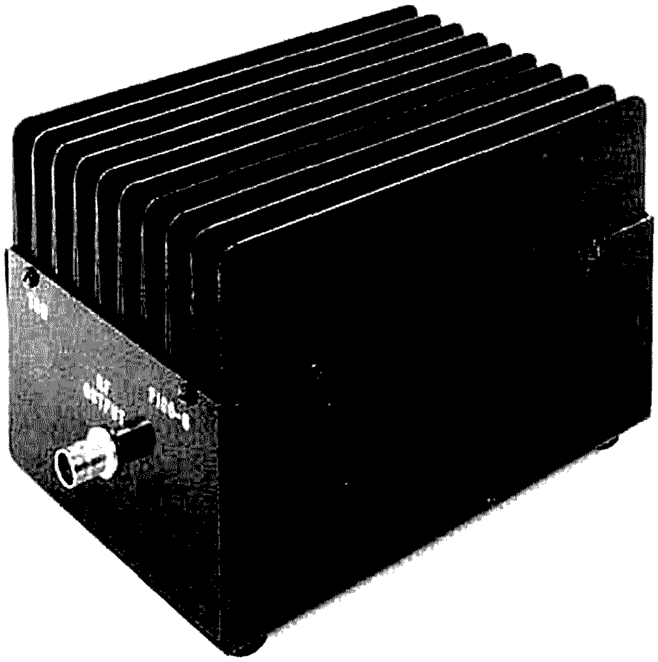
In case you haven't noticed, solid-state technology has finally caught up to the vacuum tube. A good example of modern semiconductors at work is illustrated in the high-power linear amplifier featured in this issue. Modern transistors also provide good gain and low noise well into the microwave region, and microwave power devices are now available. If you think the prices are still too high, forget for a moment those 2C39s, 2K25s and other exotic hardware you bought for peanuts on the surplus market. Compared with the decade-old *list* price of vacuum tubes that would do the same job, modern high-performance semiconductors are a best buy.

While millions of research hours (and dollars) have been spent in the pursuit of higher gain, lower noise and higher power rf and microwave devices, even more has been spent in the field of digital logic. First it was the low-frequency RTL logic family that got all the attention, then DTL, TTL, Schottky TTL, ECL and cmos – each family offering more speed or less power consumption than the last. Now IBM researchers have developed a silicon structure that creates a vacuum-tube triode in silicon. Well, not exactly – but the new device has the same space-charge-limited current flow that occurs between the cathode and the plate in a vacuum tube. The value of this achievement is that it is now possible to build very low-power logic that operates in the 10- to 100-MHz region.

This new logic, which is called SCL (for space-charge-limited), outperforms all other logic, power-wise, at switching rates over 1 MHz. Cmos circuits, while low-power kings at the lower frequencies, require more power than SCL devices at frequencies above 1 MHz. There is also a good possibility that these new SCL devices will be very attractive for low-level linear amplifiers. When operated at starved collector current levels of less than one microampere, SCL devices have shown current gains as high as 100,000. Furthermore, SCL devices theoretically should have all the low-noise performance of vacuum tubes because they have the same built-in noise cancellation that comes with space-charge-limited current flow. SCL semiconductors, of course, will be free of the heater noise that makes building low-noise vacuum tubes such a problem.

In an SCL device the space-charge-limited current flow takes place in the silicon substrate *under* a conventional lateral transistor which is located on the surface of the chip. The emitter, base and collector, in addition to providing connections to the device, provide the biases which form the operating fields that turn the n-type substrate into an SCL. The positive bias between the surface emitter and collector provide the cathode-plate potential (although only a fraction of a volt as compared to hundreds of volts in a vacuum tube, the principle is the same). The bias on the surface base creates a grid that controls the current between the cathode and plate. When the base is unbiased, a deep depletion region extends down into the silicon chip, virtually cutting off current flow. When the base is forward biased, the depletion region shrinks, allowing current to flow. The surface transistor, while notoriously slow, is never biased completely on, so it does not affect the speed of the SCL.

Jim Fisk, W1DTY
editor-in-chief



high-power solid-state linear power amplifier

Complete construction details for a broadband linear with 320 watts output — including details on combining four basic amplifiers to build a solid-state kilowatt

Steve Chambers, Applications Engineer, TRW

Chances are, when you think of building a high-power, high-frequency linear rf power amplifier, you immediately think of using vacuum tubes. There has long been a need for high powered, solid-state rf amplifiers that offer highly reliable broadband operation, reducing the need for regular preventative maintenance. However, the achievement of such solid-state linear power amplifier (LPA) designs has been inhibited by the limitations of available transistors. Until recently they had low output power levels, required elaborate temperature compensation schemes and required precise power output control during conditions of high load vswr.

Several months ago, however, TRW Semiconductors introduced a new transistor developed for linear high-frequency ssb operation that is tolerant of mismatch, overdrive and wide temperature variations. This device is the TRW PT6665A/PT5788, rated at 100 watts,

either peak envelope power or CW. The PT6665A is in a flange-mounted package while the PT5788 is the stud-mounted version. These devices, in small quantities, cost about $36.00 each. When you consider the simplicity of the design and the lack of expensive high-voltage components (and a high-voltage power supply), this is quite reasonable. However, you do have to provide a rather husky 28-volt dc supply.

This article shows how to build a 320-watt (output) linear power amplifier using four TRW PT6665As in a two push-pull pair configuration. The broadband amplifier operates directly from a 28-volt dc source and covers the frequency range from 1.5 to 30 MHz without tuning. Four of these basic power amplifiers can be combined through summing networks, as will be discussed later, to build a conservative 1000-watt linear. Although it is possible to obtain 1000-watts *output* with only three of the basic amplifiers, IMD performance is reduced slightly, particularly at the upper power levels.

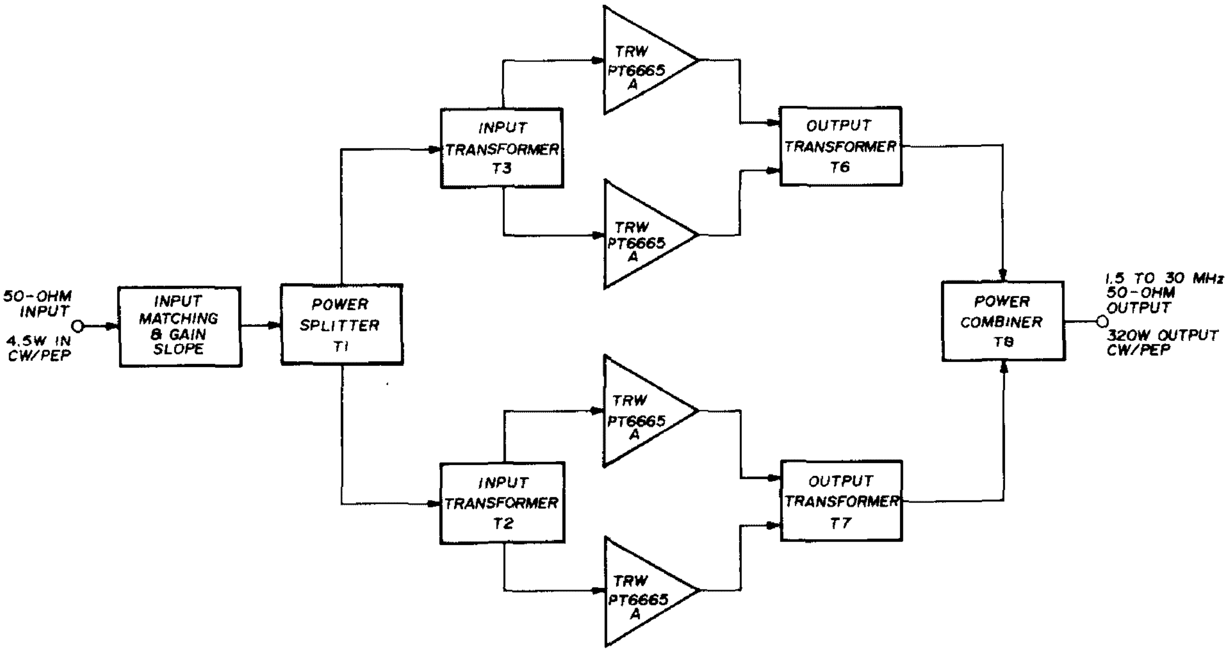


fig. 1. Block diagram for the broadband solid-state 320-watt rf power amplifier that covers the frequency range from 1.5 to 30 MHz.

The 320-watt linear amplifier shown in block form in fig. 1 has a power gain of about 17 dB. As can be seen from fig. 2, 4.5 watts of drive power is all that is required for full power output at 30 MHz; less than two watts of drive is required for full rated output on 160 meters. This amplifier is capable of withstanding open- and short-circuit load conditions at full power output and the intermodulation distortion (IMD) is better than −32 dB. If power output is held to 250 watts, the IMD performance is better than −35 dB as plotted in fig. 3.* This is better than many vacuum

*IMD referenced to either of two equal tones as is standard amateur practice. IMD must be increased 6 dB for reference to peak power, or increased 3 dB for average power reference.

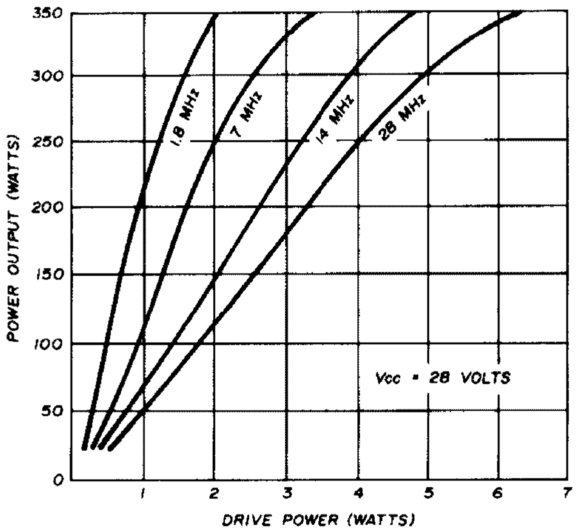


fig. 2. Drive requirements for the 320-watt power amplifier.

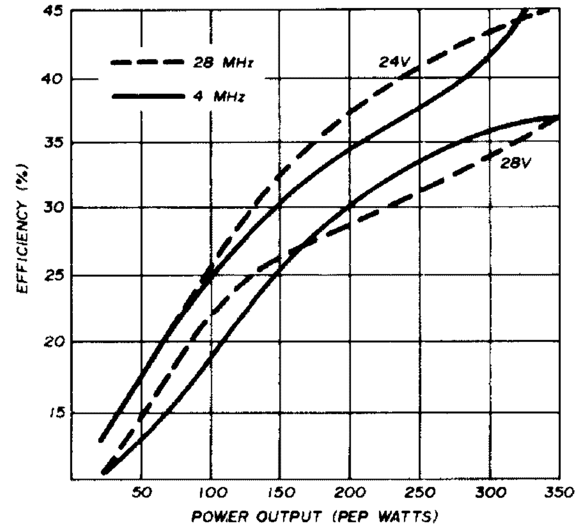


fig. 5. Gain and efficiency of the 320-watt linear amplifier vs supply voltage.

tubes, particularly the TV sweep tubes that are often used in amateur service at this power level. The affect of quiescent (idling) collector current upon IMD is shown graphically in fig. 4. Gain and efficiency of the amplifier are shown in fig. 5.

circuit

Since class-B or -AB linear amplifiers are linear only with regard to their power-transfer characteristics, the output signal contains harmonics that are a function of the ratio of the cutoff frequency to the operating frequency and to the selectivity of the output matching network. This indicates that the power transistors be operated push-pull. With this arrangement, 40-dB rejection of the even-order harmonics is readily achieved, and the odd-order harmonics can be easily filtered. To achieve the goal of 320 watts *output* in this amplifier, two pairs of TRW PT6665A transistors are operated in push-pull and their outputs are combined in a zero-degree hybrid transformer (T8) as shown in fig. 6.

The input drive to the amplifier is divided equally between the two push-pull stages by the power splitter, T1, another zero-degree hybrid transformer. These transformers convert the 50-ohm source and load impedances into two 100-ohm parts which are in phase. Any

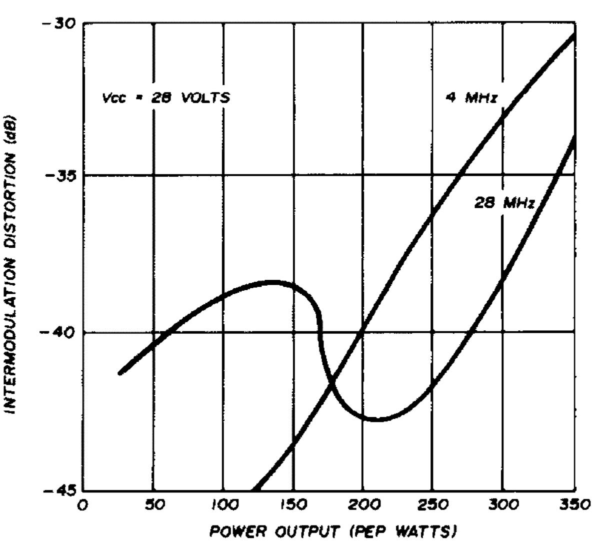


fig. 3. Intermodulation distortion of the solid-state linear amplifier at various power levels.

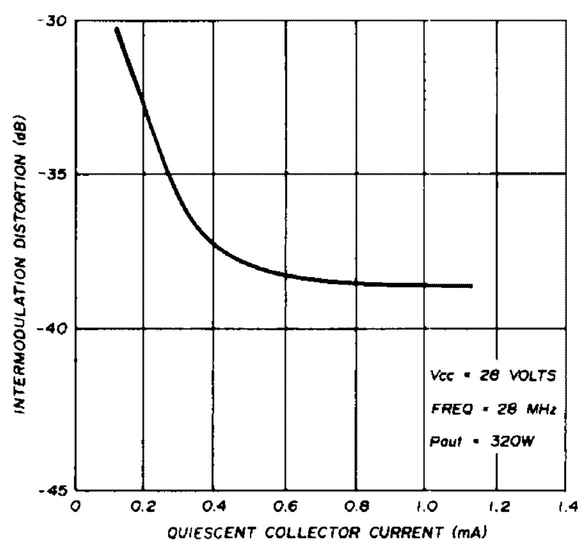


fig. 4. Affect of quiescent (idling) collector current upon intermodulation distortion.

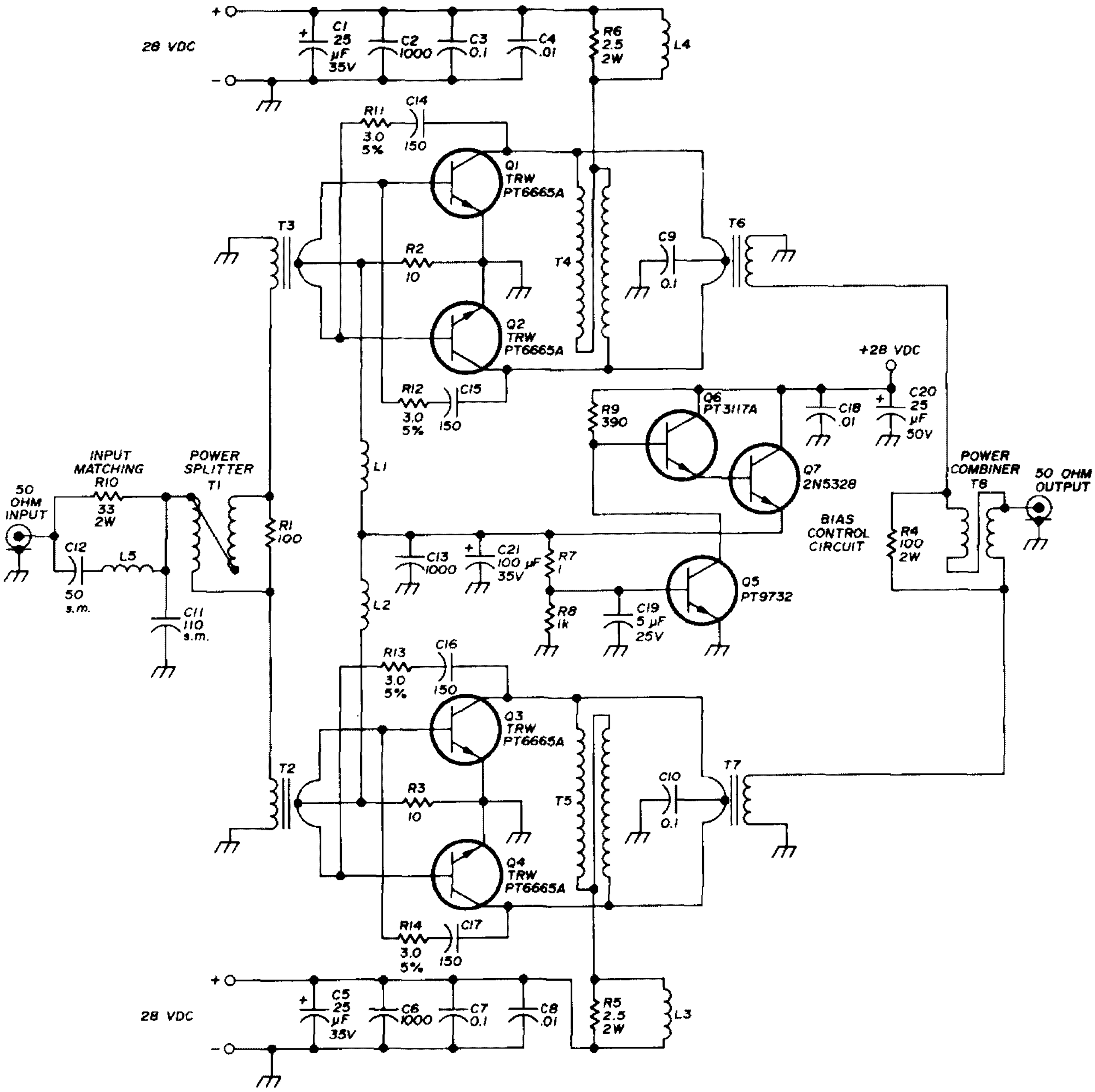
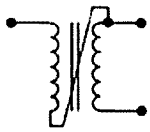


fig. 6. Schematic diagram of the solid-state 320-watt linear power amplifier. Complete construction details for the transformers are shown in figs. 7 and 8. All resistors are ½-watt unless otherwise noted.

C1, C5	25-µF, 35-volt electrolytic
C2, C6, C13	1000 pF metal clad (Underwood Electric type J-101)
C14, C15, C16, C17	150-pF metal clad (Underwood Electric type J-101)
C19	5-µF, 25-volt electrolytic
C20	25-µF, 50-volt electrolytic
C21	100-µF, 35-volt electrolytic
L1, L2, L3, L4	5 turns no. 20 enameled on one Fair-Rite CN20 ferrite bead (available from Amidon Associates)
L5	0.56-µH molded inductor
Q1, Q2, Q3, Q4	TRW type PT6665A
Q5	TRW PT9732 (thermally connected to heatsink)
Q6	TRW PT3117A (200 mA transistor)
Q7	TRW 2N5328 (2 amp transistor)
R5, R6	2.5 ohms (four 10-ohm, ½-watt resistors in parallel)

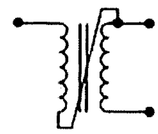
amplitude or phase unbalance causes power to be dumped into resistors R1 and R4. The input impedance of the amplifier is near 50 ohms on 160 through 15 meters, going up slightly, to 75 ohms, on 10 meters:

1.8 MHz	50 + j3.5 ohms
3.5 MHz	49 + j1.0 ohms
7.0 MHz	48 + j1.3 ohms
14.0 MHz	49 + j3.2 ohms
21.0 MHz	55 + j10 ohms
28.0 MHz	70 + j2.5 ohms
30.0 MHz	75 + j0 ohms



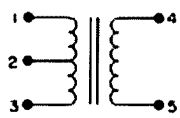
T1

Twisted pair no. 18 wire, 5 twists per inch (not critical), wound through two rows of CN-20 ferrite beads, 3 beads per row.



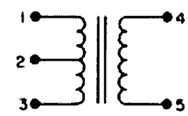
T4, T5, T8

Twisted pair no. 18 wire, 5 twists per inch (not critical), wound through two rows of CN-20 ferrite beads, 12 beads per row.



T2, T3

Terminals 1 and 3: One turn consisting of two pieces brass tubing, 0.190" (5 mm) OD, 8.80" (20 mm) long, each piece of tubing threaded through 3 CN-20 ferrite beads. Terminals 4 and 5: 3 turns no. 18 enamelled wire, wound through centers of brass tubing.

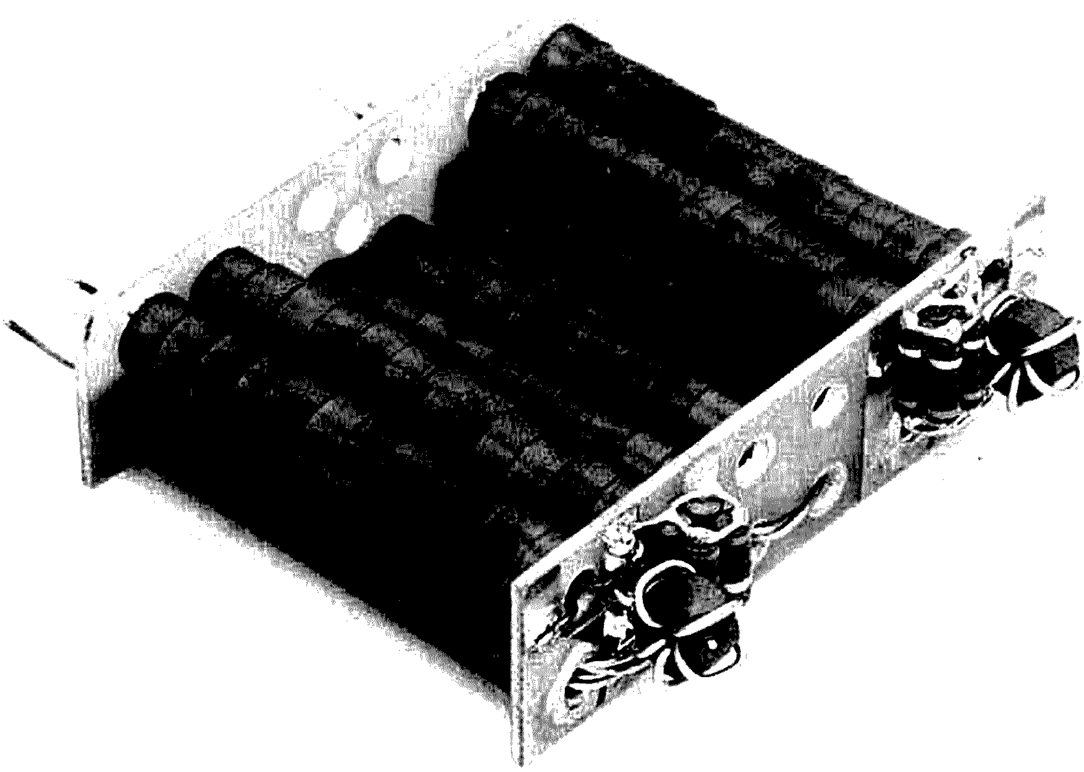


T6, T7

Terminals 2 and 3: One turn consisting of two pieces brass tubing, 0.190" (5 mm) OD, 2.50" (64 mm) long, each piece of tubing threaded through 12 CN-20 ferrite beads. Terminals 4 and 5: 4 turns no. 18 enamelled wire, wound through centers of brass tubing.

fig. 7. Winding details for transformers T1 through T8. All ferrite beads are Fair-Rite type CN-20 (No. 2643002401), length = 0.190" (5 mm), OD = 0.380" (10 mm), ID = 0.190" (5 mm). The completed transformer assembly is shown in fig. 8. The CN-20 beads are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607; 144 beads, $13.00.

fig. 8. Construction of the combiner transformers T4, T5, T6, T7 and T8. Transformer T8 is wound through two rows of ferrite beads in center; T5 is wound through two rows of beads and is mounted above T7 (left), transformer T4 is mounted above T6 (right). Complete winding details are given in fig. 7. Inductors L4 and L5 are to the right next to resistors R5 and R6 (see fig. 11).



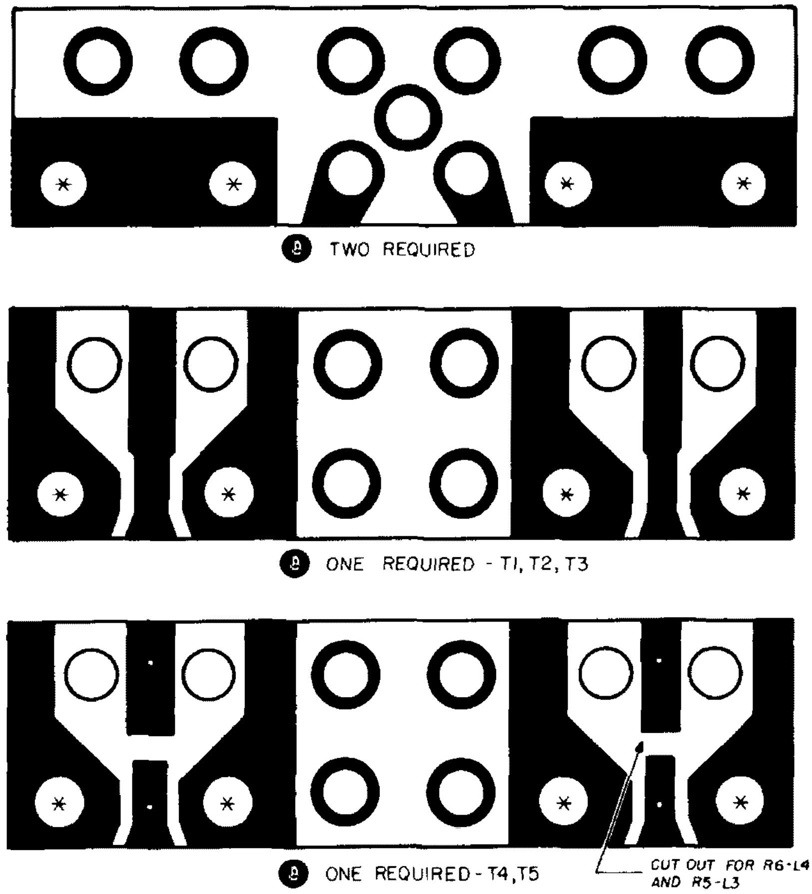


fig. 9. Printed-circuit board for the combiner transformers uses single copper-clad board. Two pairs of boards are required for the amplifier. The 0.190" (5 mm) OD brass tubing is soldered in the holes marked with an asterisk.

Transformers T2, T3, T6 and T7 each employ two ferrite-loaded brass tubes which form a center-tapped, U-shaped winding. The high-impedance winding is threaded, in continuous turns, through the brass tubing until the desired turns ratio is achieved. Winding and construction details for these transformers are shown in **fig. 7** and 8.

The required turns ratio for the input transformer, T1, is determined by:

$$\frac{N1}{N2} = \sqrt{\frac{Z_{in}}{Z_{nom}}}$$

where Z_{in} = summing port impedance (100 ohms)

$$Z_{nom} = \sqrt{Z_{LF}\, Z_{HF}}$$

The quantities Z_{LF} and Z_{HF} are the complex input impedance of the transistors at the low- and high-frequency extremes, respectively. For the TRW *PT6665A/PT5788, these values are:*

1.5 MHz: $Z_{LF} = 8.1 - j8 \cong |11.38|$ ohms
30 MHz: $Z_{HF} = 2.0 + j2 \cong |\ 2.83|$ ohms

Therefore, Z_{nom} = 5.67 ohms and the required turns ratio for the input transformer is:

$$\frac{N1}{N2} = \sqrt{\frac{100}{5.67}} = 4.2$$

The turns ratio for the collector transformer is determined by the following equation:

$$\frac{N1}{N2} = \sqrt{\frac{Z_L\, P_o}{2(V_{cc} - V_{sat})^2}}$$

where: Z_L = summing port impedance (100 ohms)
P_o = combined output power for the pair of transistors (200 watts)
V_{cc} = collector supply voltage (28 volts)
V_{sat} = rf saturation voltage (1.5 volts)

For this amplifier,

$$\frac{N1'}{N2'} = \sqrt{\frac{100 \times 200}{2(28 - 1.5)^2}} = 3.8$$

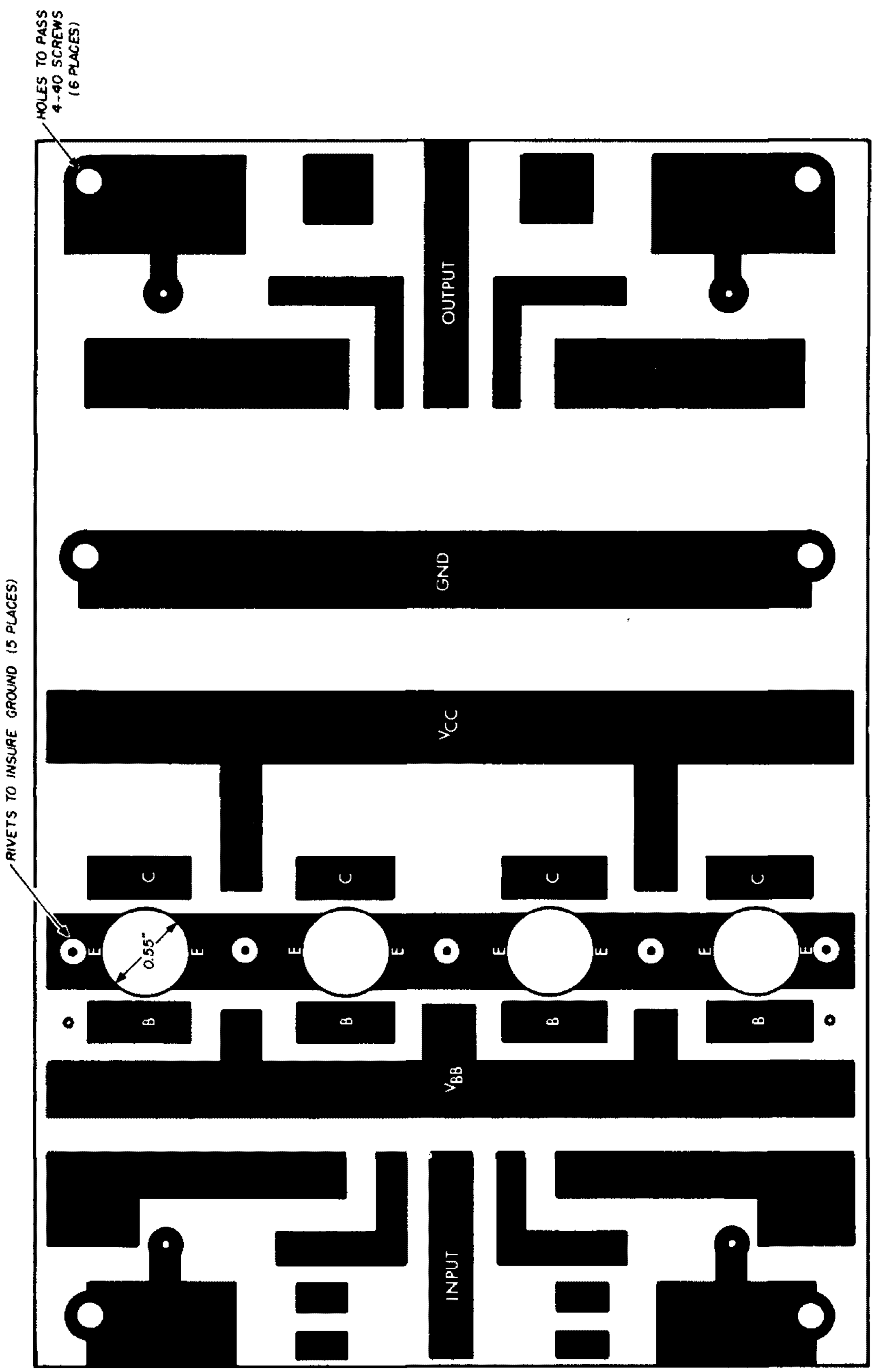


fig. 10. Full-size printed-circuit layout for the 320-watt power amplifier. Parts placement is shown in figs. 11 and 12.

The output transformer chosen for the amplifier uses the calculated turns ratio. However, for the input transformer a 3:1 ratio was used in place of the calculated 4.2:1 because it improves the match at 28 MHz. The gain-vs-frequency response and

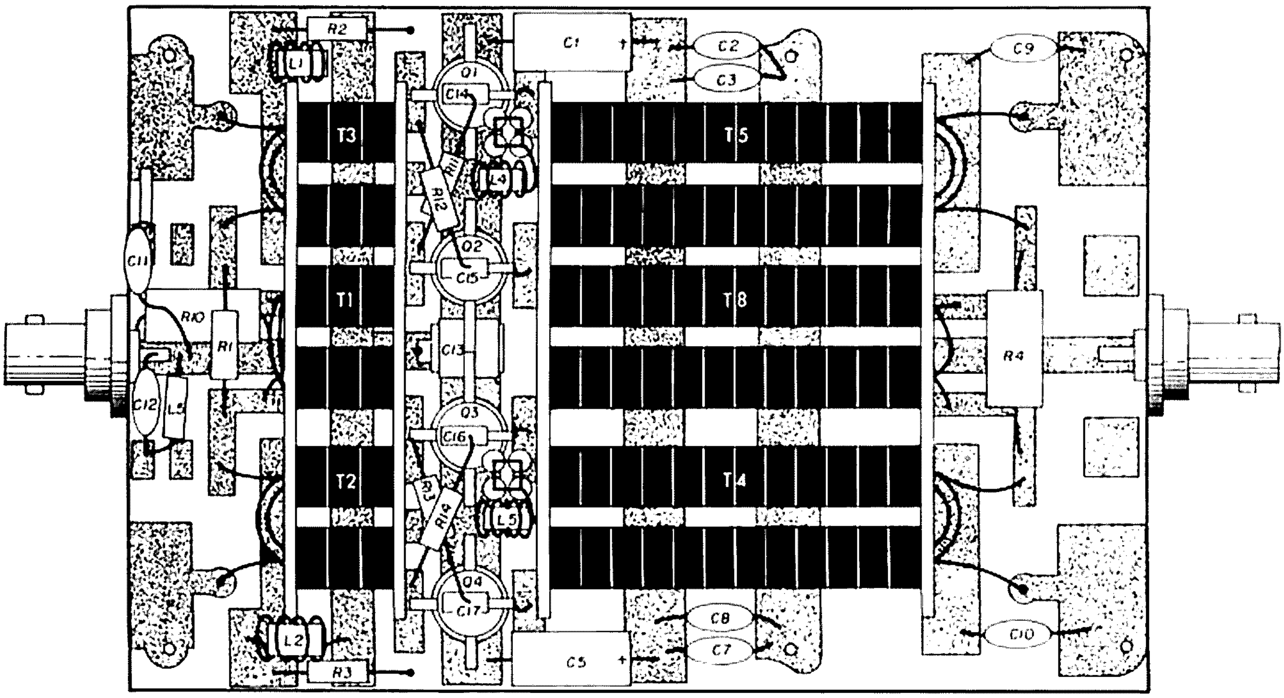
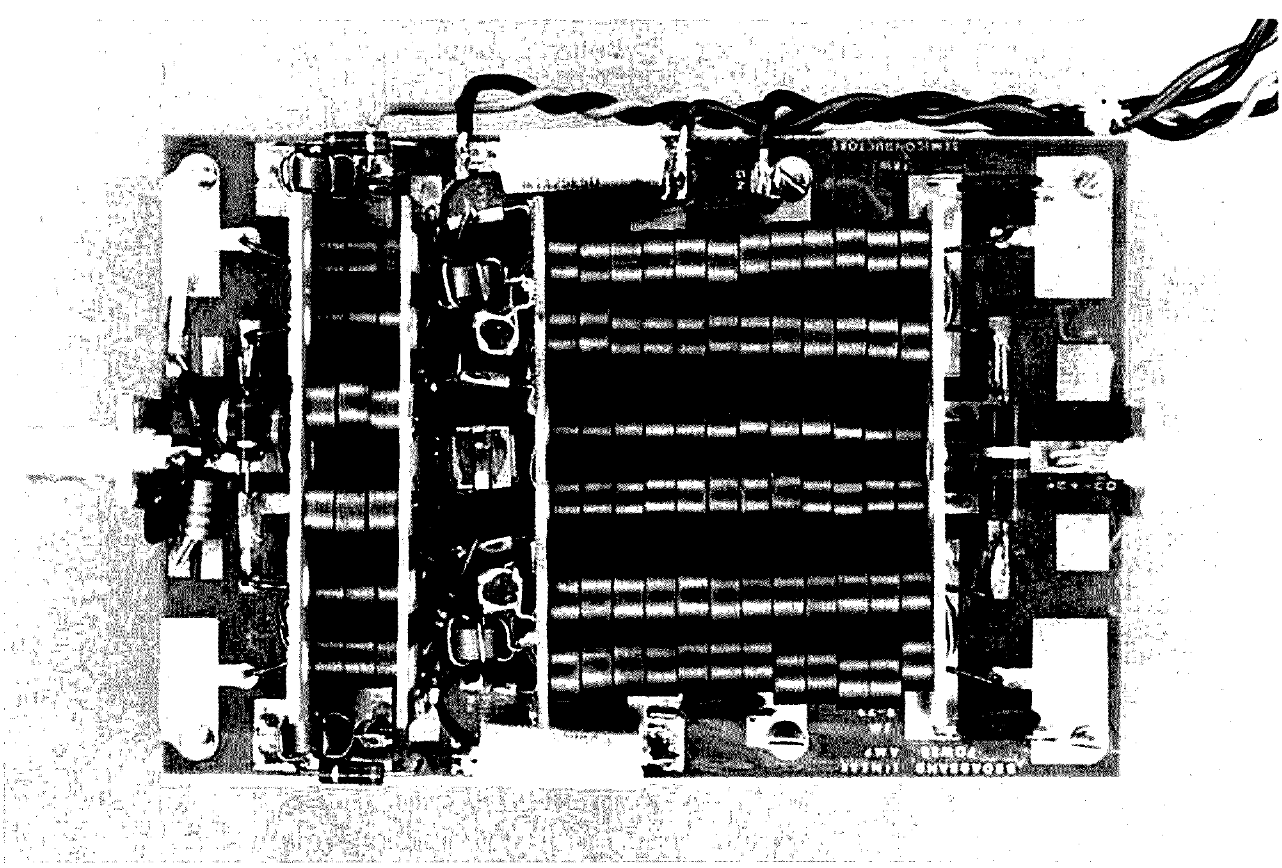


fig. 11. Component layout for the rf power amplifier. Transformer T6 is underneath T4, T7 is underneath T5.

the input impedance match have been further tailored by the addition of C11, C12, L5 and R10 at the amplifier input.

The collector feed transformers (T4 and T5) combine with the output matching transformers to form a modified 180-degree hybrid combiner as described by Pitzalis and others.[1] The ferrite mate-

fig. 12. Completed 320-watt linear power amplifier, showing location of the power transistors, combiner transformers and other components. In this photograph the input is to the left, output is to the right (see fig. 11).



rial used must have an initial permeability of 800 and the permeability must remain above 200 at 30 MHz. Losses in this ferrite are quite low and the ferrite temperature used is typically less than 20°C at full CW output. The Curie temperature is 150°C minimum (165°C typical).

When winding the transformers, care must be taken to avoid scraping insulation from the wire. Any burrs should be removed from inside the brass tubing and a heavy varnished Formvar-type wire should be used. Do not use thermal strip-away wire because it may break down and short out under high-power rf loads.

The bias control circuit used in the amplifier is of the temperature-tracking, fixed-current type (transistors Q5, Q6 and Q7). The temperature sensing transistor, Q5 (TRW PT9732), is mounted on the heatsink as close as possible to the center of the mounting area of the rf power transistors Q1, Q2, Q3 and Q4.

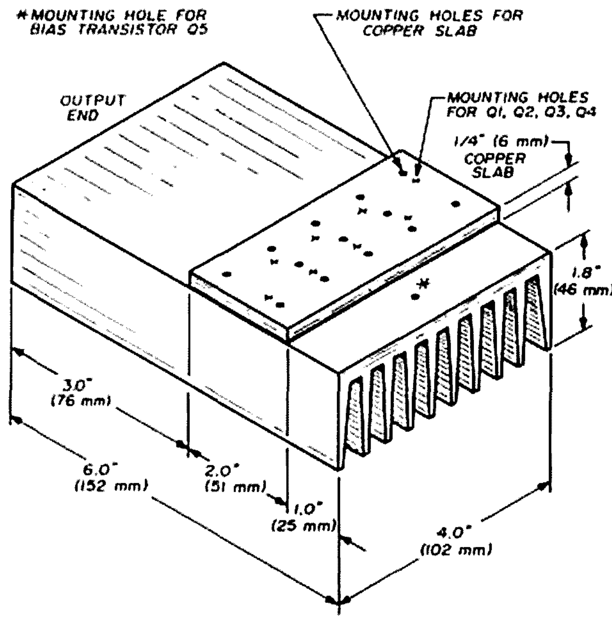
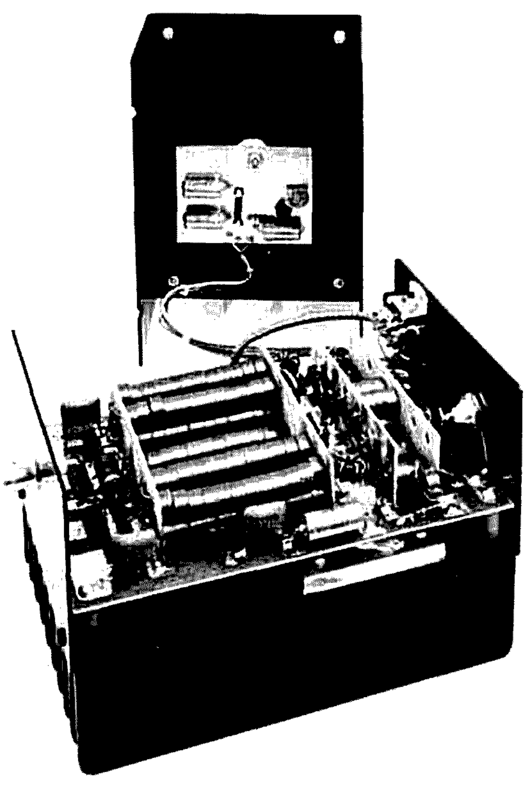


fig. 13. Heatsink and copper slab hole patterns. The heatsink is a Thermalloy 6157 or similar. All holes in the copper are drilled to pass a 4-40 screw; all holes in the aluminum are drilled and tapped for 4-40 screws. The printed-circuit mounting holes should be located after the PC board is drilled.

construction

The printed-circuit board for the 320-watt linear amplifier is shown in fig. 10. Parts placement is shown in fig. 11. This drawing also contains information on the location of the power supply filter



Complete 320-watt solid-state linear showing printed-circuit board installed on heatsink. Circuitry on small board mounted on cover is part of the bias-control circuit.

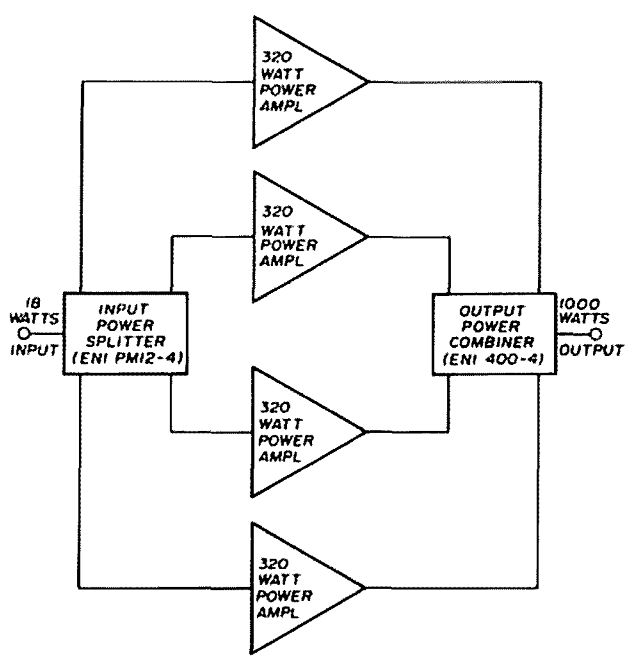


fig. 14. Four of the 320-watt amplifiers can be combined, as shown here, to provide 1000 watts output over the frequency range from 1.5 to 30 MHz. IMD and drive power for this arrangement are plotted in figs. 15 and 16.

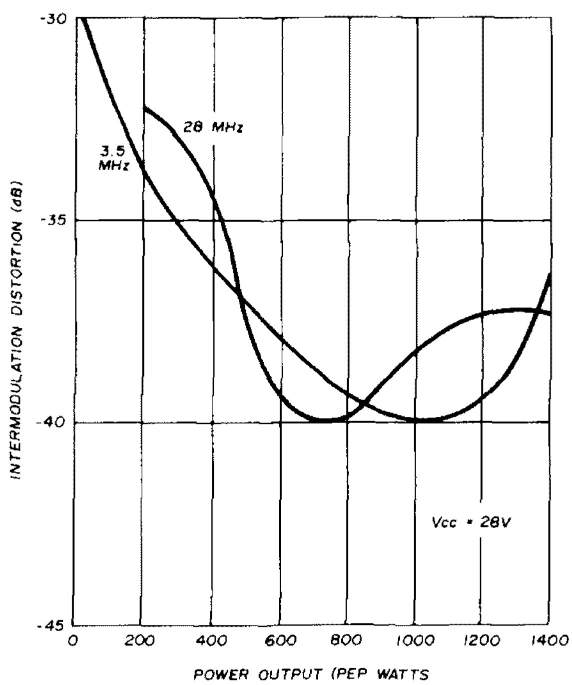


fig. 15. IMD performance of the 1000-watt linear amplifier.

components L4/R6 and L3/R5. The location of the collector-to-base feedback networks–C14/R11, C15/R12, C16/R13, C17/R14–can be easily determined from the photographs. Capacitors C2 and C6 (not visible in **fig. 12**) are located on the ground strip. Capacitor C2 is between Q1 and Q2, and C6 is between Q3 and Q4.

Since considerable heat is dissipated by the power transistors, good thermal conductivity between the transistors and the heatsink must be insured. This is accomplished by:

1. Placing a piece of slab copper between the transistor flange and the aluminum heatsink (see **fig. 13**).

2. Sanding smooth the copper slab and the heatsink surface as well as the bottom of the transistor flange.

3. Using thermal conductive compound between the copper slab and heatsink and between the transistor flanges and the copper.

4. Using cooling air for long transmission periods.

*Available from Electronic Navigation Industries, Inc., 3000 Winton Road South, Rochester, New York 14623.

solid-state kilowatt

To obtain a very conservative 1000-watts rf *output* over the frequency range from 1.5 to 30 MHz, four of the 320-watt amplifiers can be combined using straightforward, commercially available summing circuitry as shown in **fig. 14**. The input power splitter, an ENI model PM12-4, and the output power combiner, an ENI model 400-4, are available from Electronic Navigation Industries.* The overall amplifier operates directly from a 28-volt dc power supply with a typical IMD of −36 to −38 dB at *full rated output*. Although IMD performance falls off at lower power levels, it is better than −30 dB for all cases (see **fig. 15**).

The drive requirements for the 1000-watt amplifier, plotted graphically in **fig. 16**, vary from approximately 4.5 watts at 3.5 MHz to 18 watts at 28 MHz. As noted previously, a 1000-watt amplifier can also be built with three of the basic 320-watt power amplifiers, but with some increase of IMD harmonics, to −30 dB or so, at the upper power levels.

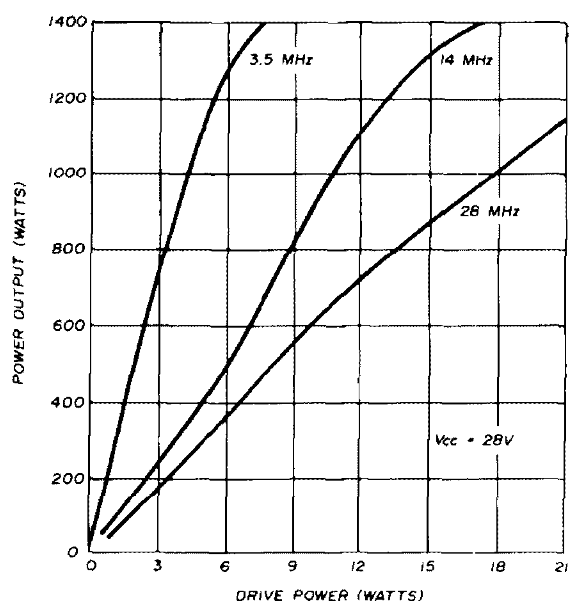


fig. 16. Drive requirements for the broadband, 1000-watt linear amplifier.

reference

1. O. Pitzalis, Jr., and T. Couse, "Broadband Transformer Design for RF Transistor Power Amplifiers," U.S. Army Electronics Command, Fort Monmouth, New Jersey

ham radio

how to calculate wind loading on towers and antenna structures

A discussion of the effect of wind loading on self-supporting towers, guyed towers and other antenna structures

John J. Nagle, K4KJ, 12330 Lawyers Road, Herndon, Virginia 22070

Almost all amateur operators are involved at one time or another in the erection of an antenna tower. The construction of a tower to withstand the elements, most notably wind, is of prime importance to the successful operation of an amateur station. Yet, very few amateurs bother to calculate the wind forces, or wind loading as it is technically known, on their antennas or towers. This may be because there is very little material on this subject available in the amateur literature.

Wind loading is generally considered to be a civil engineering subject and therefore not suitable for an electronic engineering publication, but when properly organized, the subject is relatively straight forward, requiring at most a knowledge of high school physics. Actually, I have seen subjects that are much more complex, both physically and mathematically, successfully treated in *ham radio* and similar magazines.

The purpose of this article is to organize and present the subject of wind loading on radio towers in such a manner that the average amateur who has the skill and ability to assemble and operate an amateur station can calculate the wind forces trying to overturn his tower. The wind loading on parabolic antennas and

the use of guy wires will also be briefly discussed.

Much of the material in this article has been taken from Electronic Industries Association (EIA) Standard RS-222-B, dated December, 1972.[1] I highly recommend that anyone planning to construct an antenna system obtain a copy of this standard as it is easy reading and contains much interesting and useful information beyond what will be presented here.

Calculating the wind loading on an antenna is a relatively simple procedure: First, determine the projected area of the tower and antenna. Then, by applying a very simple formula the area can be converted into a horizontal force for any wind velocity. For a free-standing (unguyed) tower which is constrained only at the base, this horizontal force develops an overturning moment which the tower and foundation must resist. The situation for a guyed tower which is constrained at both top and bottom is slightly different in that tensions in the guy wires must also be calculated; these will also be considered. These calculations are all very simple and will be discussed using examples.

projected area

For the tower and beam example, I will use a forty-foot (12.2-meter) tower made of four ten-foot (3-meter) sections with a 20-meter beam on the top; this is typical of installations used by amateurs. A typical 10-foot (3-meter) section is shown in **fig. 1**; as can be seen, the tower has a triangular cross-section with six sets of cross braces per side. The main structural members are at the corners and are composed of 1¼-inch (3.2-cm) OD steel tubing; the cross-bracing consists of 3/8-inch (10-mm) OD rod, each rod 12-inches (30.5-cm) long.

The projected area of each corner leg is therefore 1¼ x 120 = 150 square inches (3.2 x 304.8 = 975.4 square cm). Since there are two legs per face

2 x 150 = 300 square inches

2 x 975.4 = 1950.8 square cm

For the cross-braces we have

0.375 x 12 x 2 = 9 square inches

0.95 x 30.5 x 2 = 58 square cm

As there are six sets of braces per ten-foot (3-meter) section

6 x 9 = 54 square inches

6 x 58 = 348 square cm

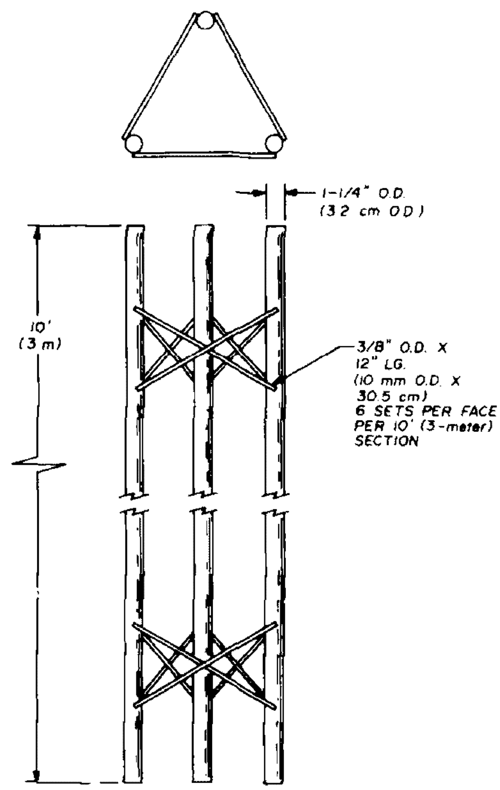


fig. 1. Dimensions of a typical 10-foot triangular tower section.

Hence, the total surface area is

300 + 54 = 354 square inches

1936 + 348 = 2284 square cm

This is equal to (354/144) = 2.46 square feet (0.228 square meters) per ten-foot (3-meter) section, as there are 144 square inches per square foot.

It is important to notice that we did not calculate the surface area of the cylindrical structural members, but instead calculated the projected area. The projected area may be defined as the outline area or as the area of a shadow cast by the member. If the structural member has a flat surface, such as a

wooden 2x4 or steel angle-iron, the projected area and surface area will be the same, but this is not the case for structural members with a cylindrical cross-section.

The reason for using the projected area can be explained as follows: When a cylindrical surface has a uniform pressure applied as shown in **fig. 2**, the components applied close to the tangential points, such as vectors A and C, exert a relatively small component on the cylinder parallel to their own direction. The radial components that they exert are equal and opposite and, hence, cancel out. The components applied at the center, such as vector B, are fully effective in creating horizontal force on the structural member. It can be shown mathematically that the wind resistance is proportional to the projected area and not the surface area.

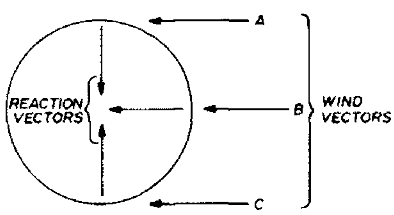


fig. 2. Diagram of wind and reaction vectors on a cylindrical surface.

In the case of a cylinder the wind force is further reduced because the streamlining effect of the cylinder makes the wind force less than for a flat surface of the same projected area. Paragraph 2.2.4 of reference 1 states that, "the pressure on cylindrical surfaces shall be computed as being 0.66 of that specified for flat surfaces." This means that when dealing with cylindrical structural members, the wind forces are two-thirds those of flat surfaces, or to say the same thing in a different manner, the projected area of the cylindrical member may be reduced to 0.66 of its actual value.

In the case of open face or lattice towers, one other factor must be considered. That is the effect of wind blowing through the tower and against the back structural elements as this area is also effective in developing a horizontal force. Paragraph 2.2.5 of reference 1 states that,

> "For open face (latticed) structures of square cross section, the wind pressure shall be applied to 1.75 times the normal projected area of all members in one face. For open face (latticed) structures of triangular cross section, the wind pressure shall be applied to 1.5 times the normal projected area of all members in one face. For closed face (solid) structures, the wind pressure shall be applied to 1.0 times the normal projected area."

The only type of solid structures I can think of that would be used by amateurs are irrigation tubing, telephone poles or wooden 2x4s bolted together.

It can be noted when dealing with lattice-type towers with a triangular cross section, and using the cylindrical structural members which are so popular in amateur work, the 1.5 triangular factor, when multiplied by the 0.66 cylindrical factor, gives 0.66 x 1.5 = 1. Therefore, both factors can be neglected when dealing with this type of tower. Although these factors may be neglected, they should not be forgotten!

wind force

So much for area involved. The problem now is to convert that area into a force. EIA Standard RS-222-B states in paragraph 2.3 that the wind pressure P in pounds per square foot is given by

$$P = 0.004\ V^2 \qquad (1)$$

where V is the wind velocity in miles per hour and 0.004 is the wind conversion factor (includes a gust factor and a drag coefficient for flat surfaces).* Note that the wind force is proportional to the wind velocity squared.

It may be pointed out that the exponent on the velocity, 2 in this case, is itself a function of the velocity. The

*In metric terms the formula is $P = 0.0075\ V^2$, where P is in kilograms per square meter and V is velocity in kilometers per hour.

factor 2 is a good average value for wind velocities in, say, the 30- to 100-mph (48- to 161-kph) region. For extremely low velocities, say, less than about 10-mph (16 kph), wind force is linear with velocity. As wind velocity approaches the trans-sonic region, the exponent becomes very high. This is why supersonic aircraft require such large engines.

It is interesting to digress for a moment and consider **eq. 1** in a different light. By applying Newton's third law (action and reaction), eq. **1** also gives the wind resistance when an object is driven at a given velocity through still air. For example, a standard size automobile presents a frontal area of about 25 square feet (2.3 square meters) so at 60 mph (96.6 kph) requires

$P = 0.004 \times 60^2 \times 25 = 360$ pounds of force

$P = 0.0075 \times 96.6^2 \times 2.3 = 161$ kilograms of force

just to overcome wind resistance. Note that increasing the velocity by $\sqrt{2} = 1.414$ (from 50 to 70 mph [80 to 113 kph], for example) will double the wind resistance. Couple this with the fact that the efficiency of a typical automobile engine is much less at 70 (113 kph) than at 50 mph (80 kph) — it's easy to see why fuel consumption increases astronomically at higher speeds.

wind-loading zones

Returning to our original problem, since we have calculated the projected area, this can be converted to a horizontal force at any given velocity by using **eq. 1**. The only question remaining is what wind velocity to design for. Fortunately, RS-222-B comes to our aid again. **Table 1** and the accompanying map of **fig. 3** give the *recommended* horizontal design wind pressures in pounds per square foot and kilograms per square meter for various parts of the United States (windloading zones) and for various heights above ground. RS-222-B also has a table giving the zones by states and counties, which will not be reproduced here because of space limitations. If you cannot pinpoint your location precisely from **fig. 3**, you may want to consult the table of counties in RS-222-B. You may also assume the more severe wind-loading zone.

table 1. Recommended horizontal design wind pressures in pounds per square foot (kilograms per square meter given in parenthesis). Wind-loading zones for the United States are shown in *fig. 3*.

height zone (above ground)	wind-loading zone A	B	C
Portion of tower 300 feet (19.4 meters) and under	30 (146.5)	40 (195.3)	50 (244.1)
Portion of tower 301 to 650 feet (91.7 to 198 meters)	35 (170.9)	48 (234.3)	60 (292.9)
Towers 651 feet (198.5 meters) and higher shall be designed for uniform wind pressure for their entire height	50 (244.1)	65 (317.3)	85 (415.0)

The data for both **table 1** and fig. **3** were obtained by statistical methods from long-term weather observations based on wind velocities that should not be exceeded, on the average, more than once every 50 years. The work is described in a paper by H.C.S. Thom.[2] A later paper on this same subject has also been published by Thom.[3] Both Thom papers are highly statistical and the EIA map offers much more usable information for the average individual.

Most amateur towers will fall in the 300-feet-and-under (91.4 meters) category for which **table 1** gives wind loading of 30, 40 or 50 pounds per square foot (146.5, 195.3 and 244.1 kg per square

meter) respectively, for wind zones A, B and C. Putting these numbers into **eq. 1** gives wind velocities of 86.6, 100 and 112 miles per hour (139.4, 161 and 180 kilometers per hour). Since I live in the metropolitan Washington, D.C., area which is clearly in wind loading zone A, I will design for wind loading of 30 pounds per square foot (146.5 kg per square meter) or 86.6 mph (139.4 kph).

practical example

Returning to the problem, we have already calculated that the projected area of a ten-foot (3-meter) section of tower is 2.46 square feet (0.228 square meter). Applying correction factors of 0.66 for cylindrical structural members and 1.5 for a triangular tower gives

$2.46 \times 0.66 \times 1.5 = 2.44$ square feet, ($0.228 \times 0.66 \times 1.5 = 0.226$ square meter) projected area

Assuming a wind velocity of 86.6 mph (139.4 kph) (or 30 pounds per square foot [146.5 kg per square meter] – gives a total horizontal force of

(2.44 square feet) x (30 lb/sq ft) = 73.2 pounds, horizontal force

(0.226 square meter) x (146.5 kg/m²) = 33.2 kg, horizontal force,

For later computational reasons we will assume this force is uniformly distributed along the length of the tower. This results in a loading of

$$\frac{73.2 \text{ lb}}{10 \text{ ft}} = 7.32 \text{ pounds per foot of tower}$$

$$\frac{33.2 \text{ kg}}{3 \text{ meters}} = 11.1 \text{ kg per meter of tower}$$

As previously stated, I intend to use four 10-foot (3-meter) sections to give the tower 40-feet (12.2-meters) height. This gives the situation shown in fig. 4 for a total horizontal force of

$$F = 40 \times 7.32 = 292.8 \text{ pounds}$$

$$F = 12.2 \times 11.1 = 135.4 \text{ kilograms}$$

Because the tower is constrained at the bottom and free at the top (unguyed), the effect of this force is to cause the

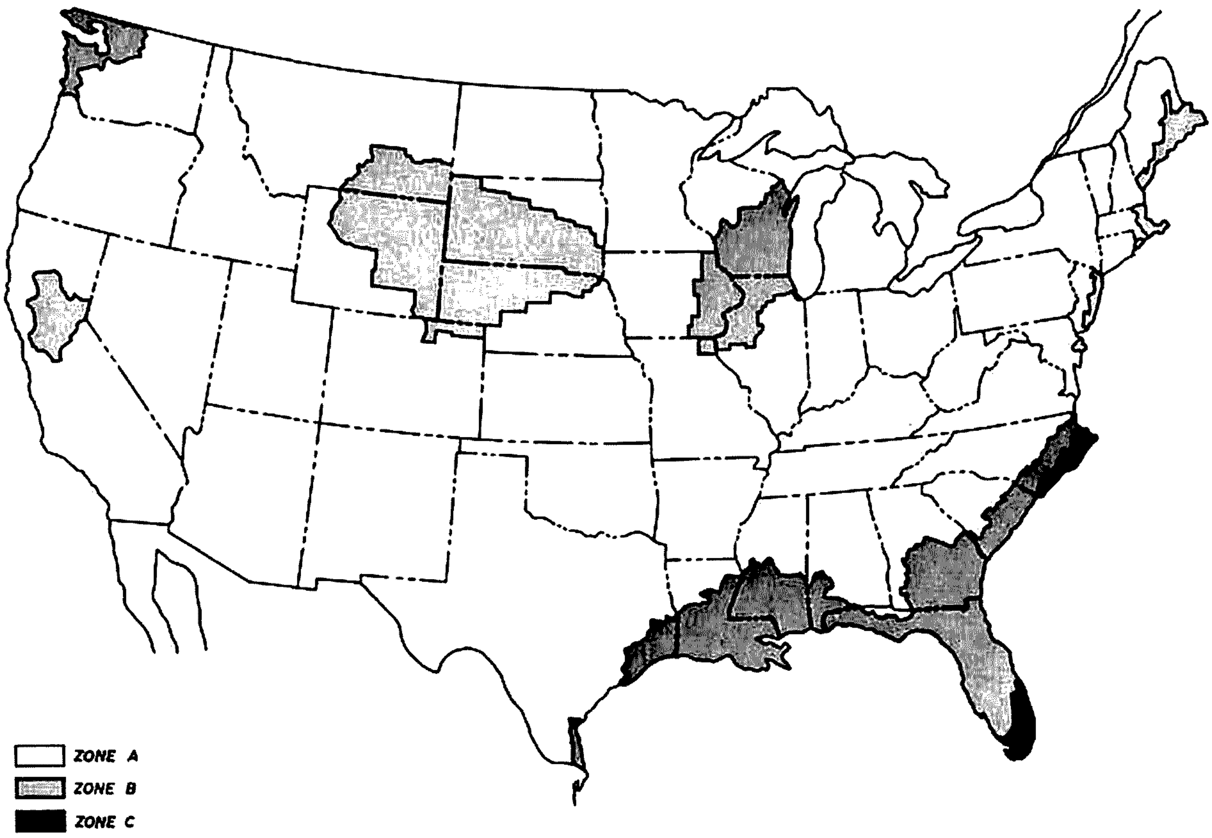


fig. 3. Location of wind-loading zones in the United States (from EIA Standard RS-222-B).

tower to rotate about the horizontal axis through its base, i.e. fall down.

In physics, a force that causes an object to rotate is called a *moment* or a torque and is defined as a force multiplied by a distance or

$$\text{moment} = \text{force} \times \text{distance} \qquad (2)$$

The units are pound-feet (dyne-cm).

The problem now is, "How do we translate a uniformly distributed horizontal force into a single overturning moment?" Luckily, the answer is quite simple; a uniformly distributed force will generate the same moment as a single force with the same total value acting at a point midway on the structure. Hence a uniformly distributed force of 7.32 pounds/foot (11.1 kg/meter) along the tower will generate the same moment as a single force of 292.8 pounds (135.4 kilograms) acting at the mid-point of the tower. This is shown in **fig. 5**. This gives an overturning moment of

$$M = 20 \times 292.8 = 5856 \text{ pound-feet}$$

$$M = 6.1 \times 135.4 = 825.9 \text{ kilogram-meters } (8.2 \times 20^{10} \text{ dyne-cm})$$

This is the moment developed by the tower alone; now let's put a HyGain Model 203-BA three-element, 20-meter beam on the top of the tower which,

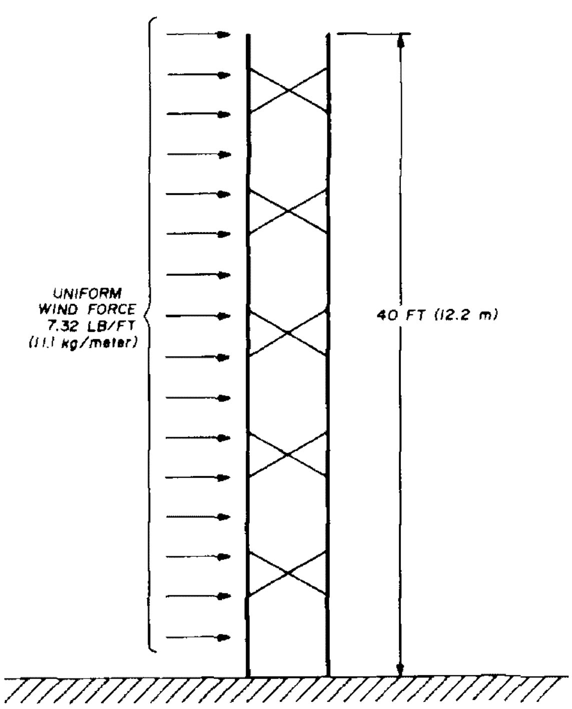


fig. 4. Uniform horizontal force on a tower.

according to the manufacturer, has an area of 3.08 square feet (0.286 square meter). A wind loading of 30 pounds per square foot (146.5 kg/m^2) (86.6 mph – 139.4 kph) will develop a horizontal force of

$$(30 \text{ lb/ft}^2) \times (3.08 \text{ ft}^2) = 92.4 \text{ pounds}$$

$$(146.5 \text{ kg/m}^2) \times (0.286 \text{ m}^2) = 41.9 \text{ kilograms}$$

on the antenna which in turn will generate a moment of

$$M = 92.4 \times 40 = 3696 \text{ pound-feet}$$

$$M = 41.9 \times 12.2 = 511.2 \text{ kilogram-meters } (5.01 \times 10^{10} \text{dyne-cm})$$

at the top of the tower. The total overturning moment acting on the tower is then

$$M = 3696 + 5856 = 9552 \text{ pound-feet}$$

$$M = 511.2 + 825.9 = 1337.1 \text{ kilogram-meters } (13.1 \times 10^{10} \text{ dyne-cm})$$

This combination is shown in **fig. 6**. This may be considered as a single force of 9552 pounds acting at a distance of one foot above the ground, a force of 9552/20 = 477.6 pounds acting 20 feet up the tower, as a force of 238.8 pounds acting at the top of the tower, or any other combination of force multiplied by distance whose product is 9552 pound-feet as shown in **fig. 6**.

In metric terms, this may be considered as a single force of 1337.1 kilograms acting at a distance of one meter above the ground, a force of 1337.1/6.1 = 219.2 kilograms operating 6.1 meters up the tower, as a force of 109.6 kilograms acting at the top of the tower, or any other combination of force multiplied by distance whose product is 1337.1 kilogram-meters.

The antenna and tower combination used in this example is relatively modest compared with some antennas, and yet, the overturning moment is nearly 5 tons on a one-foot arm!

The tower must be strong enough to transmit this moment to the foundation; the foundation, in turn, must be designed

to have a moment of its own that, when combined with the soil resistance, will resist this overturning moment with an acceptable margin of safety if the tower is to remain standing.

The design of tower structures themselves would take us into the subjects of structural mechanics and strength of materials; foundation design would add soil mechanics. These subjects are all well beyond the scope of this article. We will, however, briefly discuss guying since guy wires can significantly reduce foundation requirements.

guying

Guyed towers have the advantage of requiring much less structural material than self-supporting towers. As an example, a self-supporting 700-foot (213.4-meter) steel tower of conventional design will weigh about 460,000 pounds (208,651 kilograms). A comparable guyed tower will weigh about 200,000 pounds (90,718 kilograms) including the weight of the guys. It should be remembered that a guyed tower must also be capable of supporting the guy wires. While very few amateurs are so fortunate, or unfortunate, as to have a 700-foot (213.4-meter) tower, the savings in material of a guyed tower is obvious.

A disadvantage of guyed towers is the additional real estate required for guy wires. In a commercial installation, the designer would compare the cost of real estate against the cost of additional steel and concrete required to make the tower self-supporting. For the amateur who must install his tower in a residential neighborhood additional considerations are aesthetics and the whims of his wife. This last consideration, unhappily, is not amenable to a rigorous engineering analysis.

Assume now that the same 40-foot (12.2-meter) antenna and tower as previously described has a set of guy wires 30-feet (9.1-meters) from the base of the tower, as shown in **fig. 7.** From the Pythagorean theorem the length of the guy wire is

$$\sqrt{40^2 + 30^2} = 50 \text{ feet (15.2 meters)}$$

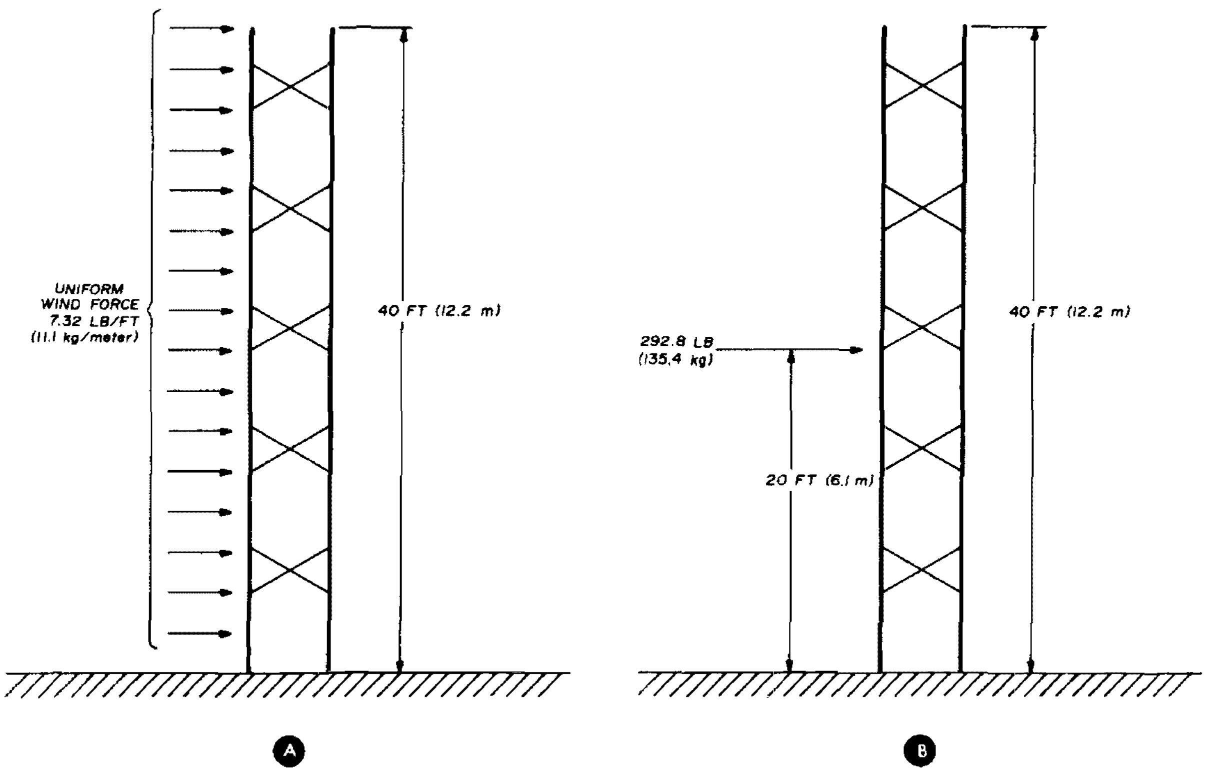


fig. 5. The overturning moment caused by the uniform wind force in (A) is exactly the same as the overturning moment generated by the single mid-tower force in (B).

The angle with respect to ground is

$$\theta = \tan \frac{40}{30} = 53.1 \text{ degrees}$$

For the sake of simplicity, we will consider only one guy wire. This is actually the case when the wind is blowing toward the tower from the direction in which the guy wire under consideration is anchored.

It is not generally appreciated, but a guy wire converts an overturning moment into increased downward stress on the tower. As previously noted, the total horizontal force in this example is 293 pounds (132.9 kilograms) on the tower plus 92 pounds (42 kilograms) on the antenna. Since the top of the tower is now constrained by the guy wires, the overturning moment is zero. For this reason the structural requirements on the tower and foundation are greatly reduced because these items need only support the weight of the assembly and the additional load imposed by tension in the guy wires.

Because the horizontal force on the tower — 293 pounds (132.9 kilograms)—is uniformly distributed over the length of the tower, we will assume that one-half the constraining force — 146.5 pounds (66.5 kg) — is located at the top of the tower and one-half at the bottom. The horizontal force on the antenna — 92 pounds (41.7 kg) — will be assumed to be located entirely at the top of the tower. This is shown in **fig. 8.**

This allocation of reactions given may seem arbitrary, but it is difficult to visualize a mechanism by which the 92-pound (41.7-kilogram) horizontal force of the antenna can be transmitted to the base of the tower without generating a moment. Also, if there is any error, the forces in the guy wire will be overestimated and not underestimated. The tension in the guy wire may be resolved into a horizontal and a vertical component. The horizontal component must be 238.5 pounds (108.2 kilograms) since it must exactly equal the horizontal windload (146.5 + 92 pounds or 66.5 + 41.7 kilograms). The tension in the guy wire itself may be calculated by setting up a vector diagram as in fig. 9.

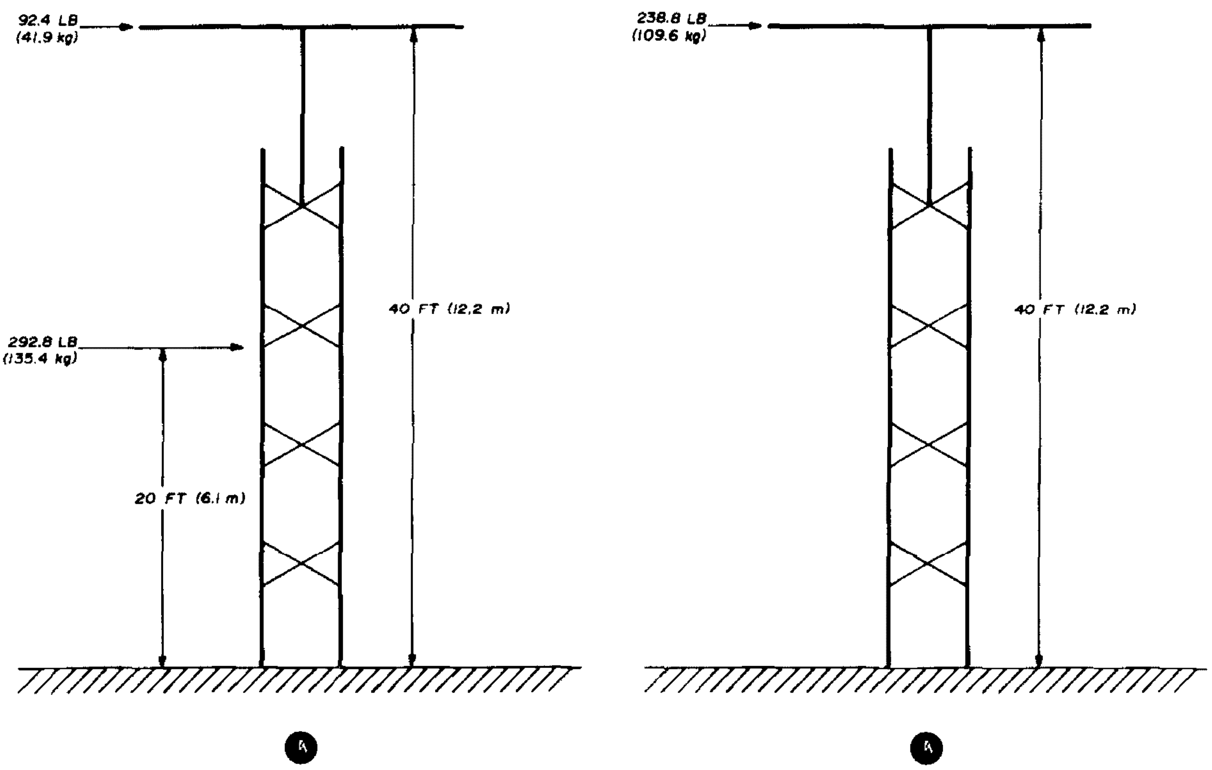


fig. 6. The overturning moment in (A) is the same as the overturning moment in (B). See text for calculations.

$$F_{gw} = \frac{238.5}{\sin 39.9^{\circ}} = 397.2 \text{ pounds}$$

$$F_{gw} = \frac{108.7}{\sin 36.9^{\circ}} = 181 \text{ kilograms}$$

where F_{gw} is the force on the guy wire and the additional vertical component which must be resisted by the tower is

$$F_v = \frac{238.5}{\tan 36.9^{\circ}} = 318 \text{ pounds}$$

$$F_v = \frac{108.2}{\tan 36.9^{\circ}} = 144.2 \text{ kilograms}$$

In other words, an 86.6-mph (139.4 kph) wind will develop a tension of 397.2 pounds (180.2 kilograms) in the guy wire and an additional vertical load of 318 pounds (144.2 kilograms) on the tower. This vertical load is more than the weight of the antenna! It should also be emphasized that these loads are in addition to the vertical loads caused by the initial tension in the guy wires.

The vertical load can be reduced by making the guy wires longer; i.e., moving the bottom end of the guy wire farther out from the base of the tower. As the guy wire approaches the horizontal, the vertical component approaches zero and guy-wire tension approaches the total wind load at the top of the tower – 238.5 pounds (108.2 kilograms).

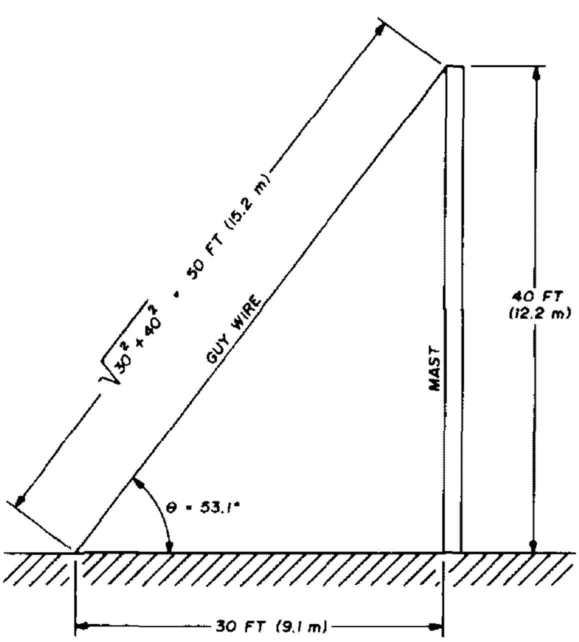


fig. 7. Guy wire arrangement. Length of guy is easily calculated by use of the Pythagorean theorem.

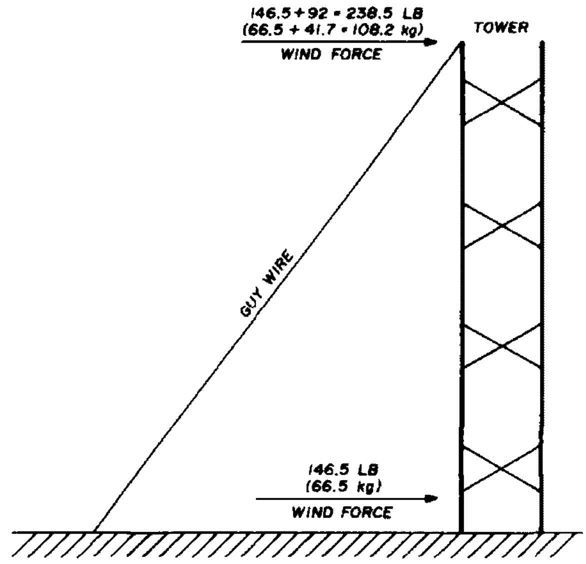


fig. 8. Horizontal forces of tower and antenna on a guyed tower.

So far we have considered the involvement of just one guy wire. When considering two guy wires you might think that for the same horizontal wind load, the tension in each guy wire would be one-half the tension for one wire. Unfortunately, it does not work out that easily. Let's consider a tower held by three guy wires at 120° intervals, as shown in fig.10. This is a typical arrangement. If the wind blows in the direction of arrow A, guy wire number-1 takes the wind load and the situation is as explained for one guy wire. If the wind blows as shown by arrow B, things are slightly different, as shown by the vector diagram in **fig. 11.** Using the same numbers as in the example above, the horizontal wind load of 238.5 pounds (108.2 kilograms) is shared equally by guy wires 1 and 2, so we allow 119.3 pounds (54.1 kilograms) on each. In this case the third guy wire carries none of the load. The horizontal force in the direction of each guy wire is thus

$$F_H = \frac{119.3}{\cos 60^{\circ}} = 238.5 \text{ pounds}$$

$$F_H = \frac{54.1}{\cos 60^{\circ}} = 108.2 \text{ kilograms}$$

As in the preceeding example, the tension on each guy wire is 397.2 pounds (180.2 kilograms) and the vertical com-

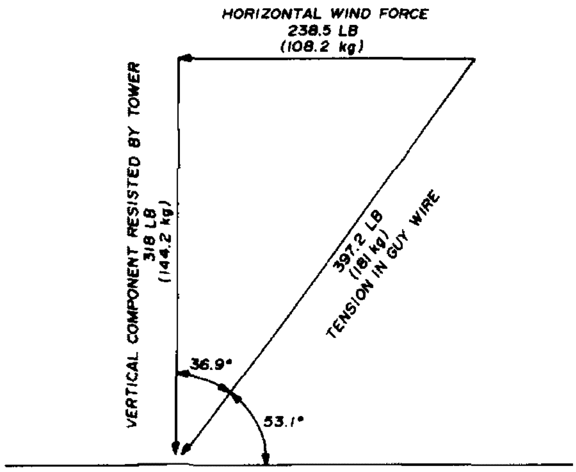


fig. 9. Vector diagram of guy wire tension. Note that the vectors represent reactions in the guy wire and not the forces acting on the guy.

ponent for each guy wire is 318 pounds (144.2 kilograms) – a total of 715.2 pounds (324.4 kilograms). The total is thus twice the vertical load generated by a wind blowing in the same direction as any one guy wire. It is easy to see that wind loads can add up pretty quickly!

The use of guy wires imposes a penalty in that the tower and foundation must be designed for greater vertical loads. The advantage of guy wires is that the tower and foundation need not be designed to resist an overturning moment. The tower designer's choice is between vertical loads and overturning moments. Also, it is much more expensive to build to resist the overturning moment than it is to withstand a straight vertical load.

An important factor to be considered in the installation of a guyed tower is the initial tension in the guy wires. If the guy wires are too loose, the tower will sway excessively. If the guys are pulled too tightly, an excessive vertical load may be put on the tower. In fact, in large installations it may be possible to buckle the tower with excessive initial tension. It is therefore necessary to compromise between stiffness and reasonably sized structural members in the tower.

antenna size

In an earlier example we assumed the area of the antenna was known from the manufacturer's data. If the antenna is homemade, I suggest that you calculate its area the same as you would for the tower. Consider each element separately *and apply the cylindrical correction factor*. Calculate the area looking down the axis of maximum radiation and also at right angles to this axis. Choose the area that is larger.

Because of the increasing interest in 1250 MHz and above, and the easy availability of parabolic antennas with solid reflectors, it is both interesting and instructive to apply the above principles to a parabolic dish and compare these with a tower. We will assume a dish with a 10-foot (3-meter) diameter and a 100 mph (161 kph) wind. From eq. 1 the wind pressure caused by a 100-mph (161 kph) wind is 40 pounds per square foot (195.3 kilograms per square meter). The projected area is

$$\frac{\pi d^2}{4} = \frac{\pi (10)^2}{4} = 78.5 \text{ square feet}$$

$$\frac{\pi d^2}{4} = \frac{\pi (3)^2}{4} = 7.3 \text{ square meters}$$

and the total horizontal wind pressure is

$$(78.5 \text{ ft}^2) \times (40 \text{ lb/ft}^2) = 3140 \text{ pounds}$$

$$(7.3 \text{ m}^2) \times (195.3 \text{ kg/m}^2) = 1426 \text{ kg}$$

If a 10-foot (3-meter) antenna is to look at the horizon to see a rising

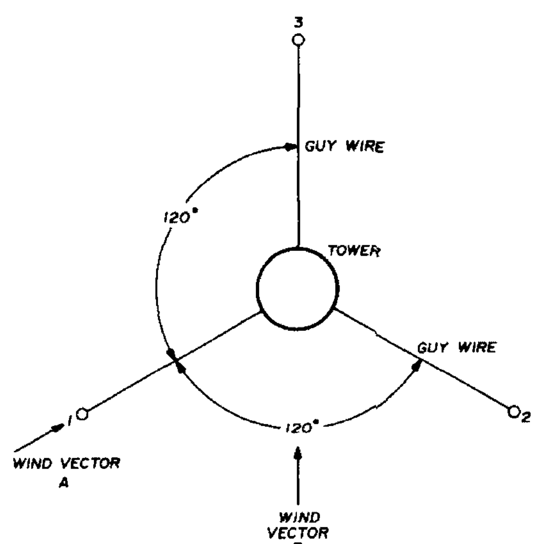


fig. 10. Tower held by three equally displaced guy wires.

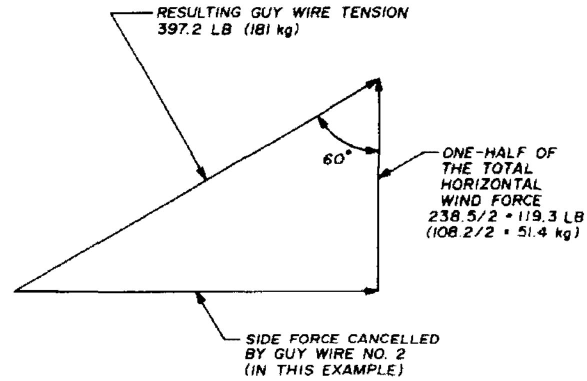


fig. 11. Vector diagram of loads in the case of two guy wires.

satellite, as shown in **fig. 12**, the mounting structure must be at least 6-feet (1.8-meters) tall to provide ground clearance. Thus, the minimum overturning moment will be

(3140 lb) x (6 ft) = 18,840 pound-feet

(1426 kg) x (1.8 m) = 2566.8 kilogram-meters (25.17×10^{10} dyne-cm)

Note that a 10-foot (3-meter) parabolic dish mounted on a 6-foot (1.8-meter) support has approximately twice the overturning moment of a 3-element, 20-meter beam mounted on a 40-foot (12.2-meter) tower.

Unfortunately, this is not the whole story. The above calculations are for a head-on wind. Experiments have shown that the greatest wind force on a parabolic antenna occurs not for a head-on condition, but when the wind is blowing at an angle to the antenna axis as shown in **fig. 13.**

For the antennas of one manufacturer[4], this wind angle is 56° and the maximum force is 10-percent higher than for a zero-degree wind angle. The reason for this is that at some wind angles the rear surface of the dish acts as an air-foil which develops *lift* in a manner similar to an aircraft wing, thereby increasing the horizontal force on the antenna.

Because this extra force depends on the shape of the antenna, it may not be the same for all parabolic antennas. The antenna parameter that has the greatest effect on this additional lifting force is probably the focal length of the antenna as this determines how deep the dish must be for a given diameter and hence will effect the lift/drag coefficient of the dish. It is not practical to give lift coefficients or angles of maximum wind loading for all cases, but a 10-percent increase in wind loading due to lift at an angle of 56° is probably a good approximation for most parabolic antennas.

The relatively large wind loading of a parabolic dish can be considerably reduced by perforating the dish. If the holes are small compared to the wavelength of operation, the effect on antenna performance will be negligible, but the wind loading will be decreased considerably.

ice loading

The preceding material has not considered the effect of ice. The magnitude of the additional load imposed by ice will depend on your location. Do you live in Miami, Florida, or Bismarck, North Dakota? The effect of ice, of course, is to increase the projected area of the structural members, thereby increasing the wind load. Unfortunately, it frequently occurs in many parts of the country that the strongest winds occur during ice storms, thereby compounding the problem. If you feel you should consider ice loads, I suggest you contact the chief engineer of a local broadcast station and find out what ice thickness his towers are designed to handle. Calculate the projected areas when loaded with the maximum expected thickness of ice. Remem-

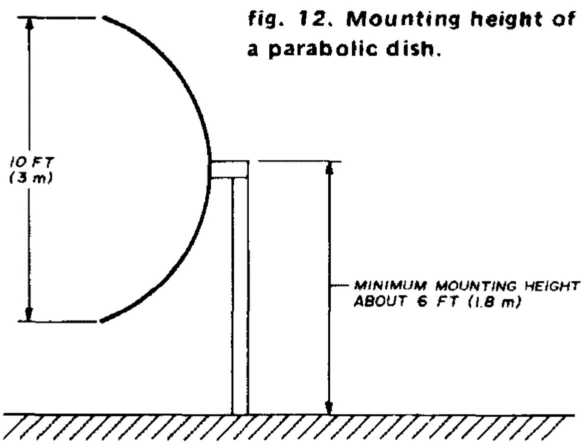


fig. 12. Mounting height of a parabolic dish.

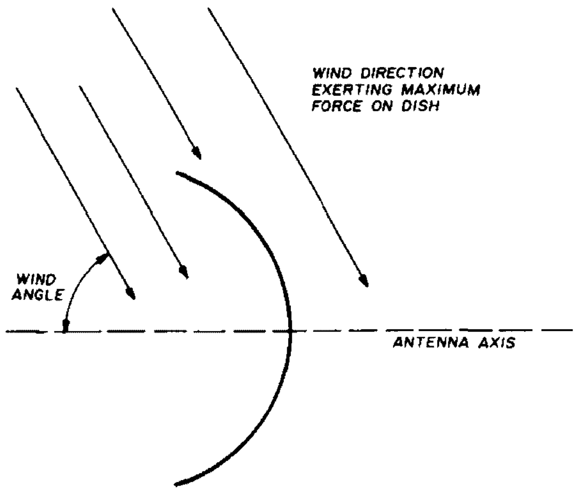


fig. 13. Relationship of parabolic antenna and direction of wind exerting maximum force.

ber, too, that ice coats both sides of the structural members so that a 1/2-inch (13 mm) radial thickness of ice will increase the overall dimensions by 1 inch (25 mm).

summary

In the above material I have shown how to calculate the overturning moments caused by wind forces on antenna structures of the type used by most amateurs. These forces can be appreciable, especially when augmented by ice. Parabolic antennas have even higher wind loadings. I have also discussed the use of guy wires, and have shown how these eliminate the overturning moment but increase the vertical loads on the tower.

The procedure for calculating the wind loading on a conventional tower/antenna combination may be summarized as follows:

1. Calculate the projected area of the tower.

2. Apply the appropriate correction factor for cylindrical surfaces and/or triangular or square cross section, as necessary.

3. Determine the maximum expected wind velocity.

4. Calculate the horizontal force on the tower.

5. Determine the effective area and horizontal force on the antenna.

6. If the tower is guyed, calculate the tension in the guy wires due to wind loading and determine the additional vertical loading of the tower.

7. If the tower is free standing, determine the moments on both the tower and antenna separately, and add.

The wind loading on a parabolic dish is calculated in the same manner as on any other structure, but maximum wind loading will occur at an angle to the main antenna beam and the numbers will be surprisingly high. Towers, like people, can carry only a limited load before they start to sway. A good man knows his own limit; a good amateur knows his tower's limit.

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2. H.C.S. Thom, "Distribution of Extreme Winds in the United States," *Proceedings of the American Society of Civil Engineers, Journal of the Structural Division,* April, 1960, page 11.
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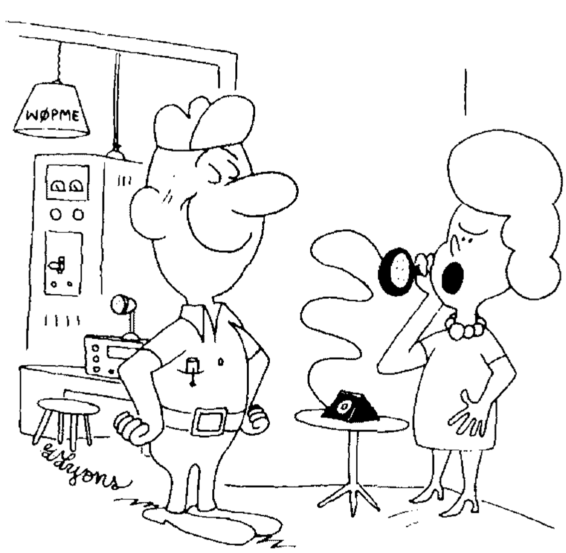
4. *Bulletin 1015B*, The Andrew Corporation, 10500 West 153rd Street, Orland Park, Illinois 60462.

ham radio



"Right now he's out of this world . . . . He just bounced a signal off of the moon!"

scanning receivers

for two-meter fm

A discussion of vhf scanner-monitors, how they work, and how they may be used on vhf fm

Joseph J. Carr, K4IPV, 225 North Glebe Road, No. 3, Arlington, Virginia 22203

For many years vhf fm monitor receivers have enjoyed a modest but reasonably steady popularity with a variety of police officers, firemen and assorted buffs, both amateur and professional. Those early models suffered from most of the same ills which blessed contemporary amateur vhf equipment: instability, difficult tuning and so forth.

The introduction of crystal-controlled, solid-state models increased the popularity of the monitor receiver within the original market but still did not cause a widespread interest among amateur radio operators. In those days, you will recall, two-meter activity was mostly a-m, scattered over a relatively large portion of the band, with no definite channelization outside of a few local net frequencies.

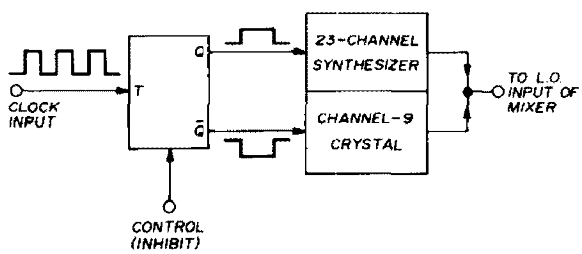
Newer monitor receivers couple crystal control with the ability to search through many channels for activity by using digital logic techniques for scanning. The scanner receiver sequentially looks at four, six, eight or even ten channels prechosen by the user. Amateur use of monitor receivers has increased due to both the availability of reasonably priced models and the channelization of our own two-meter fm activity.

simple scanner

Dual-receive Citizens Band equipment is neither vhf-fm nor amateur but is sufficiently simple and representative enough to warrant our consideration—if only for instructional purposes!

CB scanners were developed so that a single unit could be used to simultaneously monitor a regular working channel and the national emergency channel. The

fig. 1. Block diagram of a typical dual receive CB scanner receiver. A J-K flip-flop alternately grounds the 23-channel synthesizer crystal bank, then the Channel-9 crystal, permitting simultaneous monitor capability.



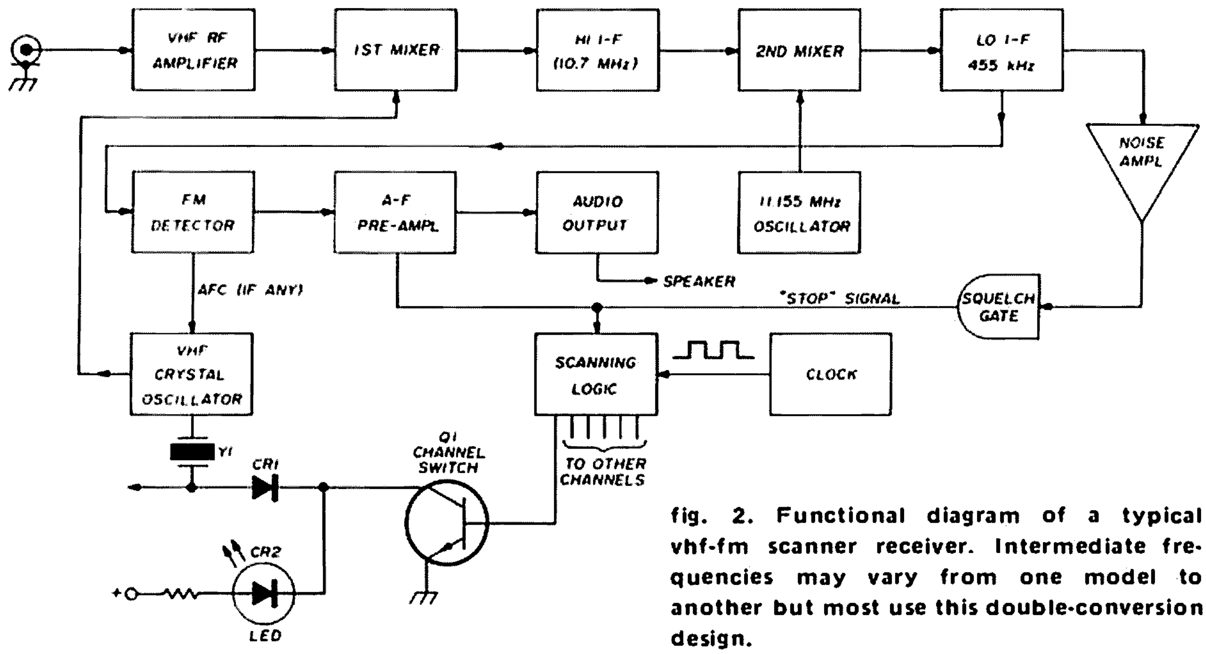
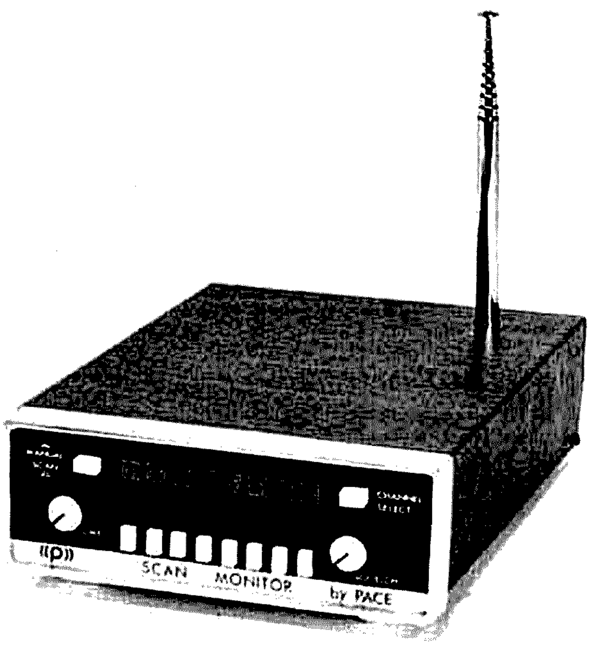


fig. 2. Functional diagram of a typical vhf-fm scanner receiver. Intermediate frequencies may vary from one model to another but most use this double-conversion design.

scanner logic switches the receiver back and forth between the channel selected by the multi-crystal synthesizer and channel 9. An increase in the agc voltage, indicating that a station is in the pass-band, causes the scanner to stop seeking and latch onto the signal. Basic operation of the circuit is shown in fig. 1. A J-K flip-flop selects which local oscillator is in control of the receiver at any instant of time. An *inhibit* signal causes the circuit to latch when the agc voltage is above a certain level.

Scan Monitor manufactured by Pace is typical of scanners used by amateurs for two-meter fm. (Photo courtesy Pace).



vhf scanners

A typical vhf fm scanner is usually a double-conversion receiver such as that shown in fig. 2. A local oscillator, operating from crystals selected by the scan logic, is connected to a mixer where it beats against the incoming rf signal to produce a high i-f in the 10- to 13-MHz range, with 10.7 MHz being most popular. A second mixer heterodynes the output of another crystal oscillator (11.155 MHz in this case) to produce a low i-f (usually 455 kHz). This signal is then handled by the receiver i-f and detector stages in the usual manner.

The squelch circuit of the monitor receiver does more than just keep the output quiet in the absence of a signal: it provides the *stop* command signal to the scanner. Without this ability scanning would be little worse than a useless nuisance.

scanning logic circuits

Fig, 3 shows the partial schematic of a scanning circuit. A unijunction transistor,

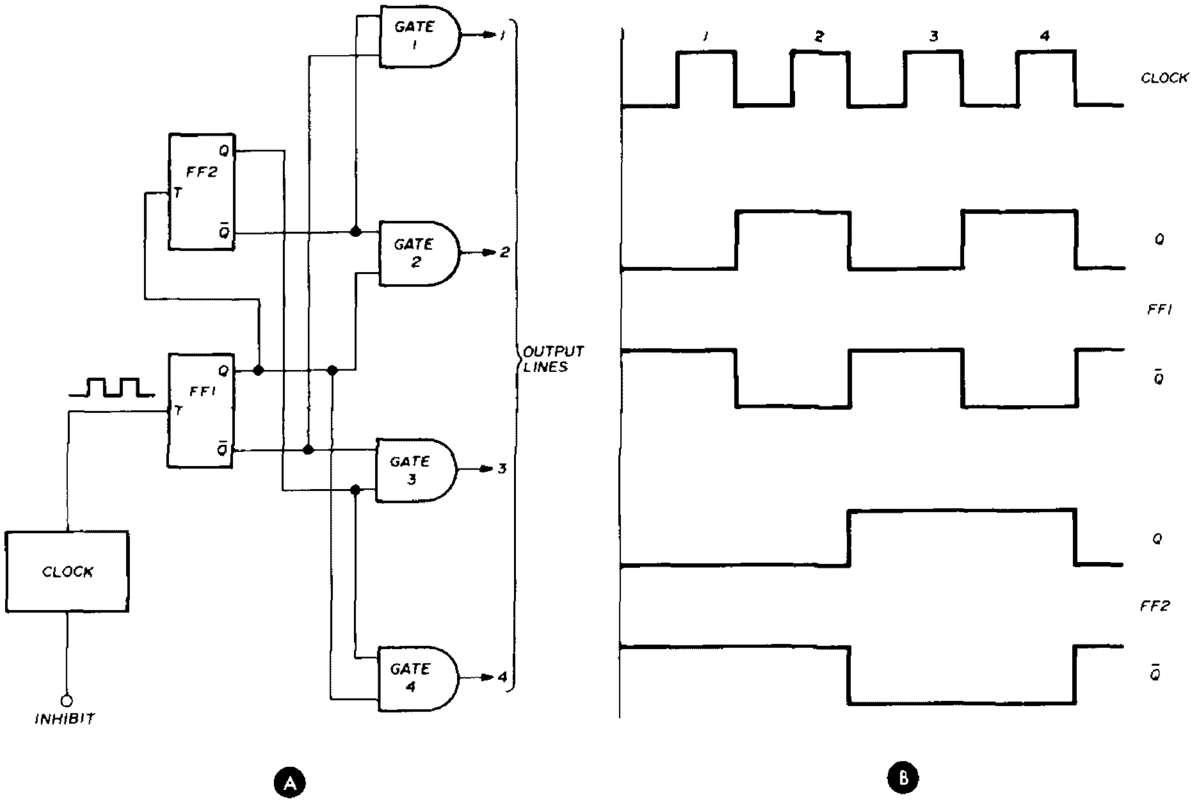


fig. 4. Logic diagram of a four-channel decoder using one dual J-K flip-flop IC and one quad two-gate (A). Waveform diagram showing design rationale for the four channel scanner is shown in (B).

Q1, operates as a pulse-generator clock. This circuit supplies sawtooth pulses to the pulse-shaping and control circuit. In that section the pulses are changed so that the counter circuits to follow see the abruptly changing waveforms they like.

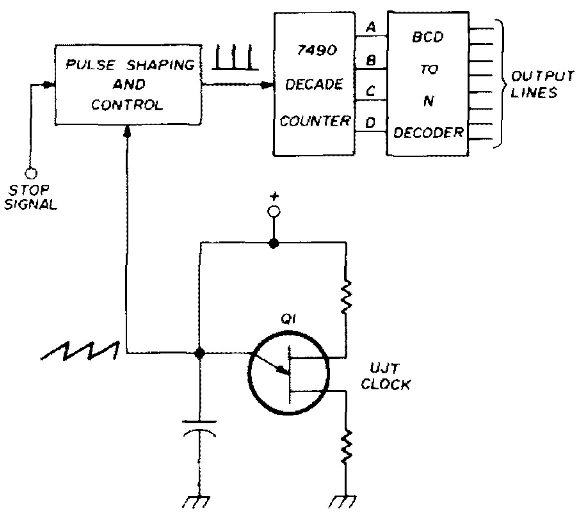


fig. 3. Partial schematic of the scanning logic of a typical receiver. In this design the BCD lines from the counter are decoded into N distinct output lines where N is the number of channels. The easy availability of IC octal and decimal decoders make eight- and ten-channel designs especially appealing.

The control portion of the circuit interrupts, on command, the flow of pulses to the SN7490 decade counter.

The Binary Coded Decimal (BCD) output from the counter is fed to a decoder which selects one of several output lines each time an input pulse is received. Examples can be found where the decoder is a suitable connection of NOR/NAND gates or an octal or decimal decoder IC such as found in decimal counting units.

The simple four-channel scanner in **fig. 4A** uses two J-K flip-flops (which are usually housed on the same IC chip) and four two-input gates (also usually a single IC). This circuit sequentially selects from among a bank of four crystals.

Waveforms which explain the operation of this circuit are shown in **fig. 4B**. The NAND gates are wired to the flip-flops in such a way that they produce a grounded output (logic zero) only when *both* inputs are high (logic 1). Notice the waveforms from FF1 and FF2 underneath clock pulse number 1. At this time only the $\overline{Q}$ of FF1 and the $\overline{Q}$ of FF2 are

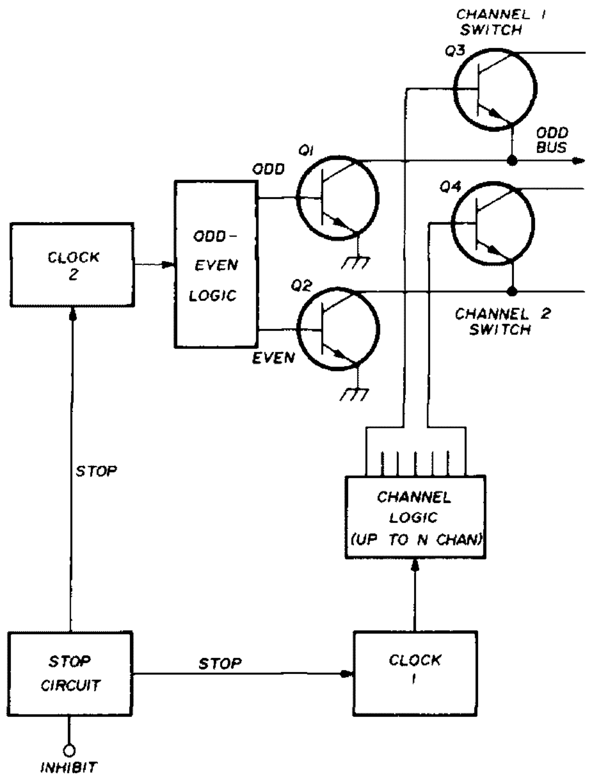


fig. 5. A simple four-channel scanner can be made to scan twice as many channels with only the added complexity of an odd-even selector. In this case the normal scan logic is the same for both channels—which is selected during any given interation of the logic signal is determined by the odd-even flip-flop.

high so they are used to drive gate 1. When clock pulse 2 arrives the Q output of FF1 and the $\overline{Q}$ of FF2 are high while all others are low. They are used to drive gate 2. This sequence continues through all four pulses from the clock, then repeats.

Most scanner receivers offer more than four channels. In fact, the standard seems to be eight. Since the binary system is based on powers of two it might be imagined that a mere doubling of the circuit of **fig. 4** would suffice. In actuality, however, the gating of eight channels is a bit more complex.

A few receivers simultaneously scan two four-channel crystal banks which are designated *odd* and *even*. One additional flip-flop sequentially selects from these two alternate banks. An example of the odd-even select system is shown in **fig. 5**.

crystal switching

Transistor Q1 in fig. 6 is the regular vhf overtone crystal oscillator used to drive the first mixer. Although the circuitry for only one channel is shown here, assume that each channel will have a similar arrangement. The cold end of the crystal, Y1, is grounded through transistor Q2 when Q2 is turned on by command from the logic circuit.

When the logic circuit selects the channel a positive voltage is applied to the base of the appropriate switching transistor. This saturates the transistor, causing a collector-emitter resistance of only a few ohms. Under this condition diode CR1 is forward biased (allowing the crystal to be grounded) and the light-emitting diode, CR2, finds a current path to ground. Depending upon design you will sometimes find lock-in or lock-out switches which will either manually select a channel or prevent it from being energized. Most scanners incorporate a small trimmer capacitor to net the crystal on their respective channels.

crystal selection

As some of us have discovered the hard way, crystals are neither absolutely calibrated (despite case markings) nor do they necessarily remain on frequency once in a circuit. The exact frequency of operation depends upon both the ambient temperature and the circuit parameters. It is, therefore, necessary to state precisely your requirements when ordering from a crystal or scanner manufac-

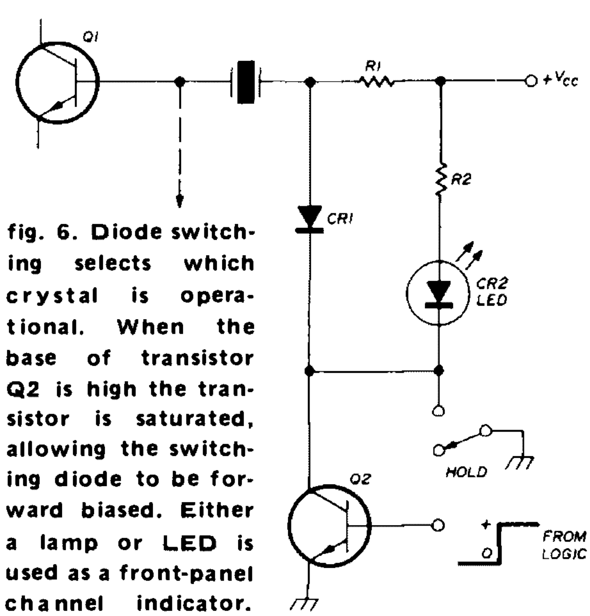


fig. 6. Diode switching selects which crystal is operational. When the base of transistor Q2 is high the transistor is saturated, allowing the switching diode to be forward biased. Either a lamp or LED is used as a front-panel channel indicator.

turer. It isn't like the old days of 40-meter CW where you bought a crystal ±2 kHz.

Crystals for any given scanner can usually be purchased from the respective dealers or from a crystal manufacturer. Before you order, especially from a crystal manufacturer, you will need certain facts about the required crystal. One piece of data, of course, is the operating frequency. To find this you must know both the channel frequency and the i-f of your unit. You also need to know whether crystal operation is in a fundamental or one of several overtone modes.

Vhf-fm scanners typically (but not universally) operate in the fundamental mode on low band (30-50 MHz), the third overtone on high band (148-174 MHz) and the ninth overtone on uhf. Assuming this to be true in your own receiver, use one of the following formulas *for determining crystal frequency:*

Low band (including 6 meters):

$$\text{Crystal frequency} = \text{Channel frequency} + \text{i-f}$$

High band (including 2 meters):

$$\text{Crystal frequency} = \frac{\text{Channel frequency} + \text{i-f}}{3}$$

uhf:

$$\text{Crystal frequency} = \frac{\text{Channel frequency} + \text{i-f}}{9}$$

Note that in some receivers the manufacturer will specify that you are to *subtract* the i-f from the channel frequency.

Prepare a simple chart for the crystal supplier listing the following:

1. Make and model of receiver.
2. Crystal frequency desired.
3. Holder style (consult catalog).
4. Mode of operation (fundamental, third overtone, etc.).
5. Circuit capacitance.
6. Drive level in milliwatts.
7. *Maximum allowable series resistance.*
8. Temperature (if in oven).

This information can usually be found in the service manual for your receiver. If a manual is not available consult Howard Sams' *Scanner-Monitor Service Data.** *This handbook covers most of the more* popular types of scanner receivers.

It is worth noting that the cost of crystals can almost double the cost of the scanner if you don't shop around a little. It is often advisable to buy a unit custom set-up from the factory with all crystals in place. *This is generally less expensive* and is also more likely to result in satisfactory performance should your local dealer be unable to provide good quality alignment service.

other scanner-receiver circuitry

For the most part the remaining

Scanner-Monitor Servicing Data*, Volume 1, SD-1, 1972, and Volume 2, SD-2, 1973, Howard W. Sams and Co., Inc., $4.95 each from Ham Radio Books, Greenville, New Hampshire 03048. Volume 1 covers the **B&K PF-1; **Browning** XM-888; **Johnson** Duo-Scan Low Range, High Range, 241-0340-001, and 241-0340-002; **Midland** 13-915 and 13-925H/L/M; **Pace** Scan 108H/L/U, 280 and 308; **Pearce-Simpson** Gladding Hi-Skan; **Penneys** 981-6065, 981-6066 and 981-6067; **Realistic** Patrolman Pro-7 (20-5001), Patrolman Pro-8 (20-162) and Patrolman Pro-9 (20-164); **Sonar** FR-104, FR-105, FR-2516, FR-2517, FR-2525, FR-2526 and FR-2528; and **Teaberry** Scan "T". Volume 2 covers **Electra** Bearcat III; **Midland** 13-922 and 13-927; **Regency** R1HT1-1, R1LT1-1, R1UT1-1, R2HT1-1, R2LT1-1, R2-UT1-1, TME-16U, TMR-1U and TMR-8U; **Tennelec** Tennetrac I/II/IV; and **Unimetrics** Digi-Scan 4+4 and Digi-Scan-8.

Also available is *Scanner-Monitor Data,* Volume III, SD-3, 1974, $5.95 from Ham Radio Books. This volume includes schematics, parts lists and service adjustments for the following scanner receivers: **Electra** Jolly Roger; **Johnson** Hi/Lo Duo-Scan, UHF/VHF Duo-Scan (late production), UHF Mono-Scan and VHF Mono-Scan; **Midland** 13-914; **Pearce-Simpson** Cherokee 8+8, Cheyenne 8 (PR-78) and Comanche 16 (PR-160); **Regency** MT-15S, TME-16H/L, TME-16H/LH/U, TME-16H/LL/U, TME-16H/LM/U, TMR-1H, TMR-1L, TMR-4H, TMR-4L, TMR-8H, TMR-8L, TMR-8H/LH, TMR-8H/LL and TMR-8H/LM.

circuits found in scanner receivers will closely parallel similar circuitry in other vhf-fm receivers including quite a few two-meter fm transceivers.

The second mixer of one popular scanner receiver is shown in **fig. 7**. Input signals to the mixer are coupled through a tank circuit and a ceramic crystal bandpass filter. The output of the mixer is tuned to the lower i-f by another, similar crystal filter.

Since most commercial transmitters use narrowband fm (±5-kHz deviation) a 12- to 16-kHz passband is required. This allows the use of many low-cost ceramic filters such as the Murata line from Japan. These same filters in different bandwidths are used in many home and auto fm and fm-stereo broadcast radios.

I-f amplification is almost universally supplied by a one- or two-stage IC amplifier. The detector might be an ordinary diode type or it might be an IC. Some scanners use the standard ratio detector/discriminator circuits which feature i-f amplification, limiting and detector diodes in one IC package. An example is the RCA CA3043. Others use an IC quadrature detector such as the Motorola MC1357.

An unusual squelch circuit is shown in **fig. 8**. This design uses switching diodes to generate squelch action. This circuit produces little audio distortion because low-level ac (i.e., audio) signals can ride on top of high levels of dc which forward biases the diode. The circuit operates by

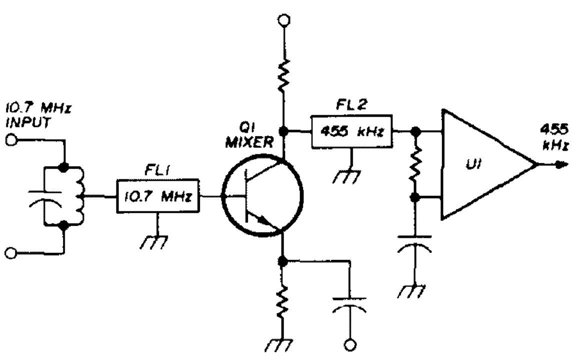


fig. 7. Most modern vhf-fm scanner receivers use modern design for the i-f strip and second mixer. This circuit uses ceramic bandpass crystal filters and an IC amplifier, U1.

causing the transistor to saturate. When that occurs the B+ to the *diodes is* shunted to ground, causing the diodes to be reverse biased. This cuts off the audio path. When the transistor is inoperative

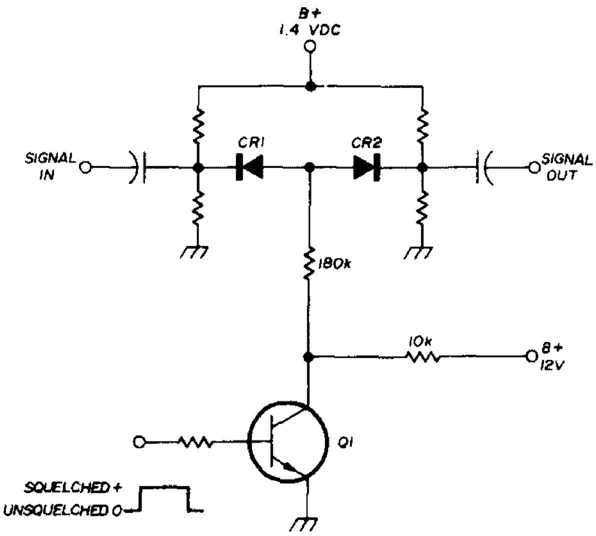


fig. 8. Diode switching is also useful in squelch circuits. When transistor Q1 is cut off B+ flows through the 10k and 180k resistors, forward biasing the switching diodes, allowing low-level audio signals to pass through. When the transistor is saturated, however, the B+ to the diodes is shunted to ground and they are effectively reverse biased.

again the B+ biases the diodes to pass audio signals.

future of scanners

The scanner market will undoubtedly be soon overrun with lower cost import and domestic models. Many of these receivers will require modification, even to the extent of completely changing the front-end tank circuits to a new range, before they can be used on the amateur bands. Others either already have sufficient range or will be available in amateur band models. In either event expect to see more amateur use of these receivers. They allow you to monitor several active frequencies or repeaters at one time. Perhaps the next logical extention of this concept is to make a transmitter in the same box which also scans. The combination could then be set to keep tabs on *all* local activity.

ham radio

integrated-circuit ssb transceiver

Complete construction details for a miniature ssb transmitter and receiver using Plessey ICs

Brian D. Comer, G3ZVC, Senior Applications Engineer, Plessey Semiconductors

This article describes the i-f and audio circuitry of a single-sideband transceiver designed by the Applications Department of Plessey Semiconductors using their SL600-series integrated circuits. The transceiver may be used at any frequency from a few kHz to 500 MHz.

The unit described in this article consists of a single printed-circuit board which requires only the addition of a local oscillator, a preselector, a linear amplifier, volume control, microphone and loudspeaker to build a complete transceiver.

receiver

The receiver is a single-conversion superhet with a 9-MHz i-f. To optimize its intermodulation performance the incoming signal is fed directly to a hot-carrier diode ring mixer and then to the crystal filter; there is no rf amplifier. The i-f sensitivity is such that at frequencies of 30 MHz or less no rf amplification is required if a reasonable antenna is used (as it would be with a transceiver). However, if the receiver is used at frequencies above 30 MHz, or with a less than ideal antenna, some rf gain may be necessary to obtain the necessary noise figure. The rf amplifier should have the lowest gain consistant with the frequency and antenna to be used and must have good large-signal handling capability if receiver performance is not to be degraded.

The mixer is an Anzac MD108 hot-carrier diode double-balanced modulator.* This device was chosen for its conveniently small size, high performance and low cost, but similar devices from other manufacturers could also be used. All the ports of this modulator are designed for 50 ohms; two have a frequency range of 5 to 500 MHz while the third covers the frequency range from dc to 500 MHz. The input from the antenna is applied to the dc- to 500-MHz port via a preselector, and the local oscillator—at a level of +7 dBm (500 mV rms)— is applied at pin 8 (see **fig. 1**). The mixer output from the rf port passes through a toroidal transformer to match it to the 500-ohm input impedance of the crystal filter. If other types of filters are used it may be necessary to re-design the impedance-matching transformer.

Once the signal has passed through the

*Anzac MD-108 double-balanced mixers are available in small quantities from Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154. The price is $7.00 each, plus postage.

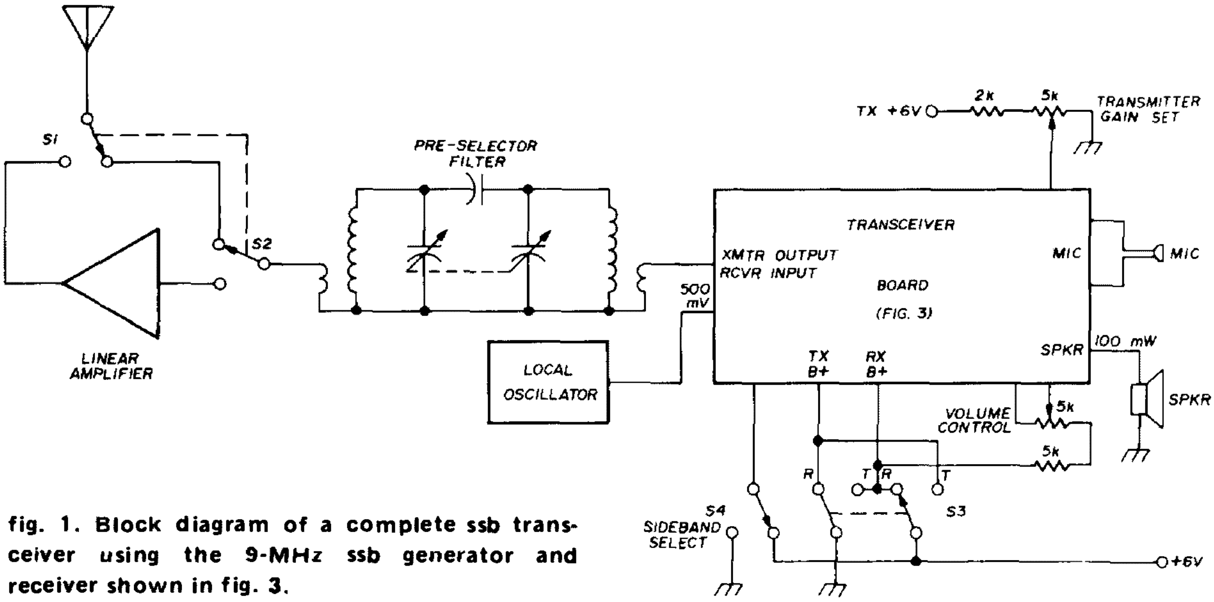


fig. 1. Block diagram of a complete ssb transceiver using the 9-MHz ssb generator and receiver shown in fig. 3.

crystal filter, a 2.4-kHz bandwidth 9-MHz filter with 90-dB stopband suppression, there is little further risk of cross-modulation or intermodulation. The i-f strip consists of three cascaded Plessey SL612C* i-f amplifiers followed by an SL640C product detector. Without agc applied each SL612C has 34-dB gain and 15 MHz bandwidth. Since a broadband i-f strip consisting of three SL612Cs has more than 100-dB gain and 15 MHz bandwidth, it *can* very easily become unstable. Therefore, the circuit board layout is very important (see fig. 4). It is relatively easy to build a three-stage, broadband i-f strip on double-sided printed-circuit board if the component side has a plane of grounded copper, but on single-sided board the layout shown in **fig. 4** should be rigidly adhered to.

The beat-frequency oscillator for the product detector uses an fet circuit that delivers about 100 mV rms to the SL640C product detector. This oscillator also supplies the carrier for the transmitter balanced modulator. One of two crystals, for upper or lower sideband, is selected by a diode switching arrangement.

The detected audio from the product detector drives an SL630C audio output stage which is capable of providing about 65 mW to headphones or a small loudspeaker. The detected audio also drives an SL621C agc system. Since the SL630C has voltage-controlled gain, the volume control consists of a potentiometer which provides a control voltage to the SL630C. If 65 mW is insufficient output (it is worth listening to it before deciding as it is usually adequate for domestic listening) an external, higher power audio amplifier may be driven either from the SL630C output or directly from the product detector.

The agc is provided by an SL621C audio-derived agc system. Its output is buffered by a transistor Q2 so an S-meter may be connected if desired. Since Q2 reduces the available agc voltage swing,

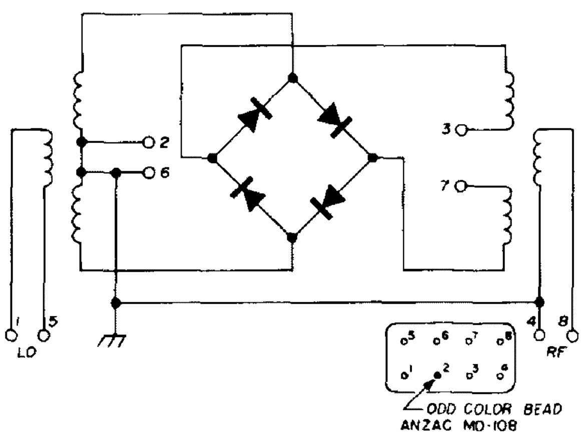
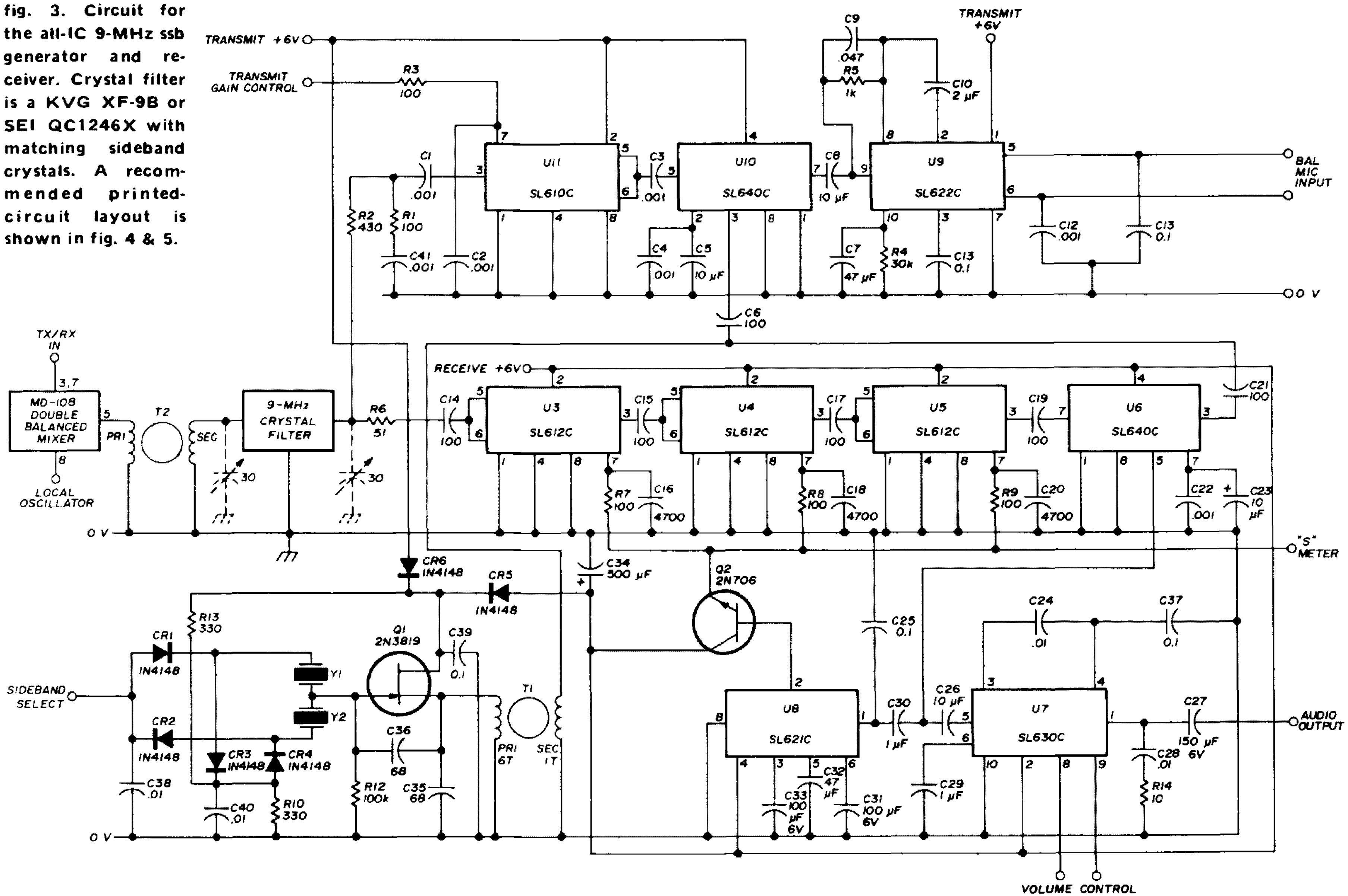


fig. 2. Schematic diagram of the Anzac MD-108 double-balanced diode mixer. Local-oscillator signal is applied to pins 1 and 5, input (output) preselector circuit is connected to pins 3 and 7, and transformer T2 is connected to pin 8.

*Plessey integrated circuits are available from Plessey Semiconductors, 1674 McGaw Avenue, Santa Ana, California 92705.

fig. 3. Circuit for the all-IC 9-MHz ssb generator and receiver. Crystal filter is a KVG XF-9B or SEI QC1246X with matching sideband crystals. A recommended printed-circuit layout is shown in fig. 4 & 5.



agc is applied to all three i-f stages to ensure that the agc can cope with the receiver's 112-dB dynamic range. If resistor R7 is replaced by a germanium diode there will be a delay to the first-stage agc—this may improve receiver noise figure very slightly on small signals—and is barely worthwhile. Capacitors C16, C18 and C20 are kept down transmitter/receiver side of the crystal filter on the change-over from receive to transmit. All the transmit/receive switching on the board is achieved by turning on the appropriate power line (transmit or receive) and grounding the unused line. The grounding of the unused line is very important as instability can result if it is not done.

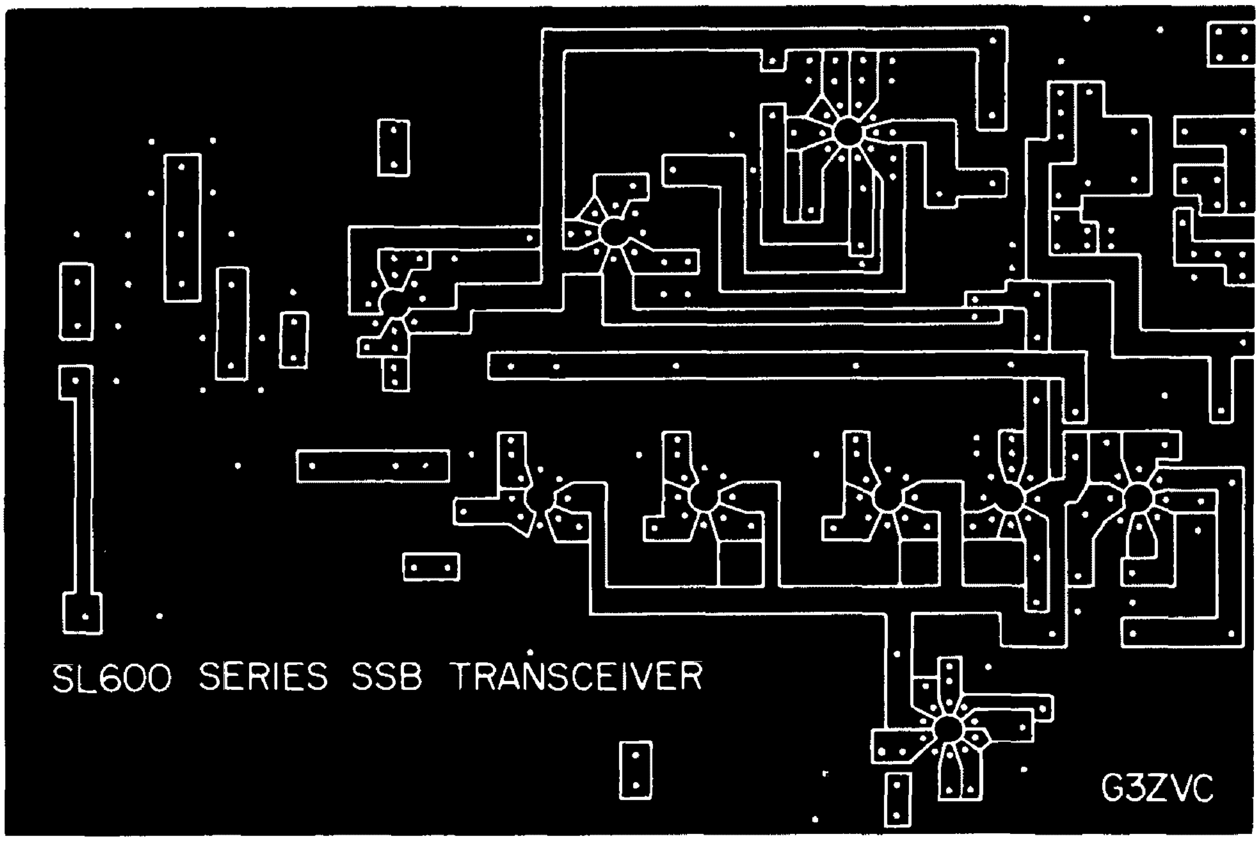


fig. 4. Full-size printed circuit layout for the 9-MHz ssb transmitter and receiver. Component layout is shown in fig. 5.

to 4700 pF to retain the ignition suppression characteristics of the system.

transmitter

The transmitter is also a single-conversion design. It generates a 9-MHz single-sideband signal using the same crystal filter as the receiver. The 9-MHz ssb is converted to the final operating frequency by the MD108 ring mixer; the unwanted frequency product is removed by the preselector. This system requires no signal switching between the antenna side of the preselector and the

The audio input from the microphone is amplified by an SL622C agc amplifier which will give a constant 100-mV rms output for a 60-dB input range. If a single-ended input is used rather than a balanced input, dynamic range is reduced to about 46 dB. In most systems 60 dB input dynamic range is too large, 40 dB being sufficient, so resistor R5 has been included in the circuit. If 60 dB dynamic range is required resistor R5 should be omitted and C9 reduced to 4700 pF.

The audio output from the SL622C

microphone amplifier goes to the SL640C double-balanced modulator. The carrier input to the balanced modulator is fed by the bfo (which works on both transmit and receive since its power may be derived from either line through diodes CR5 and CR6). The output from the impedance-matching transformer and is mixed with the local oscillator signal to provide the final transmitter frequency (and an image which is removed by the preselector). This is amplified by the linear amplifier and transmitted. The output from the preselector is about 70 mV rms.

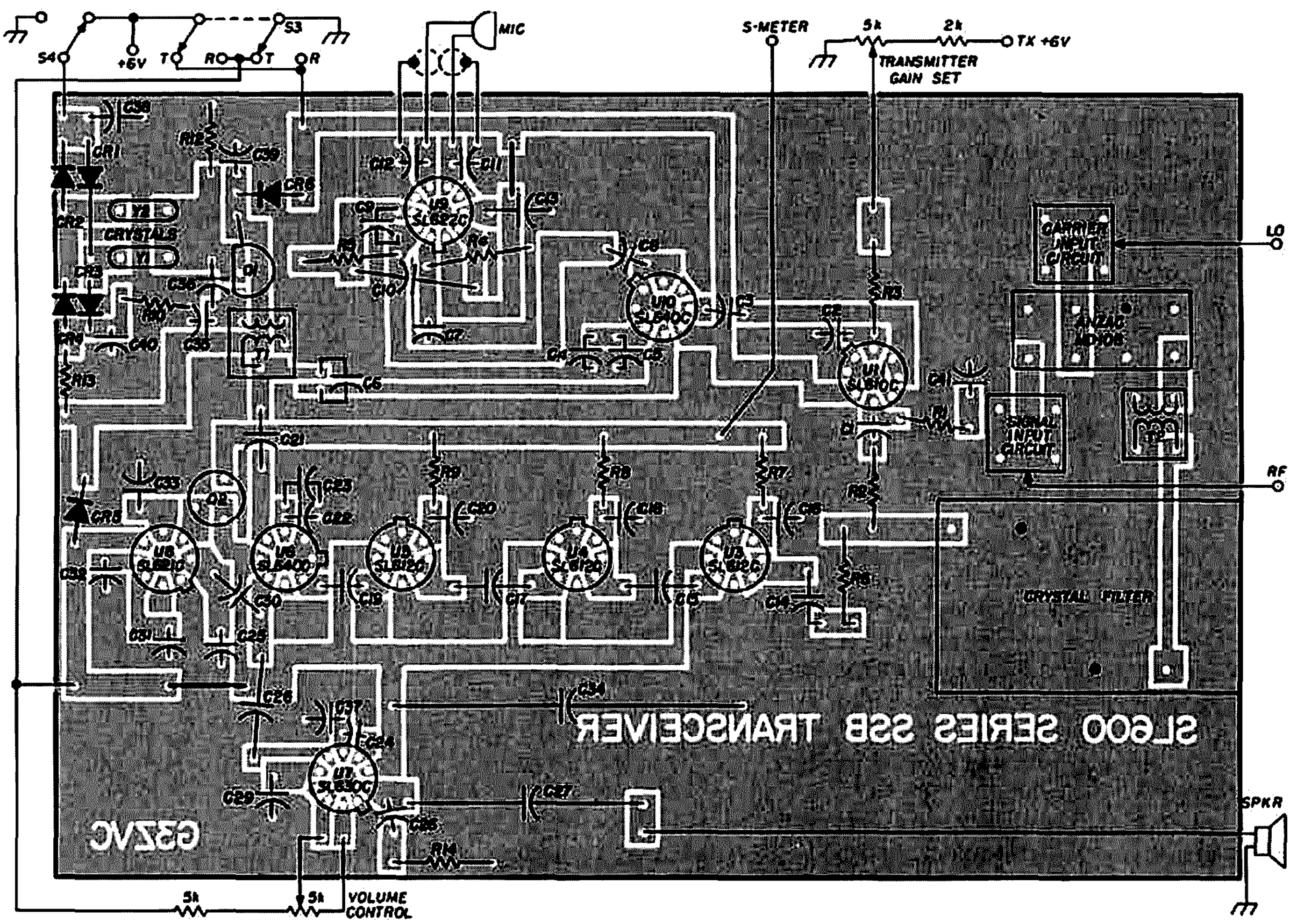


fig. 5. Component layout for the 9-MHz ssb transmitter and receiver. Space is provided for installing the input circuits for the local oscillator and signal. Design of these circuits depends upon the frequencies selected for your own application.

SL640C is a double-sideband signal with low carrier feedthrough (usually -40 dB) which is amplified by an SL610C. The gain of this particular device may be controlled either by an alc signal, derived from the transmitter linear amplifier or manually with a dc gain control. The amplified dsb signal is sent through the crystal filter to remove one sideband. Resistors R1 and R2 ensure a correct match to the crystal filter both on transmit and receive.*

The ssb output from the filter passes to the doubly-balanced diode mixer via the

construction

The complete system is built on a single-sided printed-circuit board that requires two jumpers—one in the receive supply, the other in the transmit supply. If only a receiver is required, components

*A 30-pF (nominal) capacitor from the input pin to ground will improve the passband ripple and a 20-pF capacitor from the output pin to ground will do the same. In practice, however, it has been found that these components make little difference (1-dB additional passband ripple). The KVG XF9B is the same as the SEI QC1246AX.

R1 to R5 inclusive, C1 to C13 inclusive, C10, and the semiconductors U9, U10, U11, CR5 and CR6 must be omitted, a wire jumper connected where CR5 was, and a 500-ohm resistor connected from the filter end of R6 to ground.

The layout of the board is quite critical and changes of printed-circuit design will almost certainly lead to instability unless double-sided board is used. The design shown may be built on double-sided board quite safely.

The components used in the original unit are given in the schematic. Bead tantalum capacitors are used where possible for their small size but since they are sometimes hard to find in high capacitances at high voltages, aluminum electrolytics have been specified in three places. The WeeCon capacitors specified may be replaced with other miniature high-K ceramic capacitors but the values of components should not be changed. The resistors are all 1/8-watt, 10% types.

Transformer T2 is made on a ITT CR-071-8A toroid core (the Amidon T-37-2 is a suitable substitute). Four 2-inch (5-cm) lengths of number-26 wire are twisted together and two turns are wound on the core with the twisted wire. The ends are then opened and three windings are connected in series for the filter winding and the fourth is used as the winding connected to the diode ring. Transformer T1 is wound on a core of the same type and has a 6-turn primary and a single turn secondary.

This ssb transceiver is probably the simplest which may be made using the Plessey SL600 series ICs, but its performance is not compromised in any way. It has a sensitivity of better than 0.5μV for 10-dB signal-to-noise ratio, it can handle signals of over 200 mV rms at the diode mixer with minimal intermodulation distortion, and the board uses less than 500 mW on transmit or receive. It has been designed so that anyone with basic technical competence but without previous experience in ssb transceiver design can build a successful ssb transceiver.

ham radio

harmonic phase detector

Circuit description of a simple harmonic phase detector that requires input signals to have a 2:1 frequency relationship

Henry S. Keen, W5TRS, Fox, Arkansas 72051

The **phase lock loop** has found many applications in modern electronics. One of the more important, to radio amateurs, is in the detection of fm signals where a voltage-controlled oscillator or vco is maintained at zero beat with the incoming fm signal. The control voltage generated for this application by a phase detector becomes the audio-frequency output signal. Suitable integrated circuits for this purpose are becoming available, and offer much convenience to the builder.

However, a problem in design and layout may appear wherein the vco signal may leak into a high-gain i-f system, possibly even causing saturation. An answer to this situation is the harmonic phase detector shown in **fig. 1**. This phase detector requires that the local vco operate at twice the intermediate frequency of the receiver. As a matter of fact, it won't even work if the two input signals are the same frequency! The circuit can be seen to represent a pair of peak-reading diode detectors of opposing polarity, with differential output.

The action of this circuit is illustrated in **figs. 2** and **3**. In **fig. 2**, where both signals cross the zero axis simultaneously, the positive and negative peaks of the

difference between the two input signals will be equal. Under these conditions the differential output signal will, of course, be zero. In **fig. 3**, however, when a phase difference exists between the cross-over points of the two input signals, this equality no longer exists, and a differential output voltage will be produced. The magnitude and polarity of the resultant output will be a function of the degree and direction of the phase difference.

In common with the reciprocating detector[1] this circuit was also devised to receive double-sideband suppressed-carrier signals as I am especially interested in the potential advantages of portable and mobile equipment in which the entire input to the final amplifier might be in the form of audio frequency power, obtained from a transistor amplifier, operating directly from the car battery.

By referring to **fig. 3** it can be seen that polarity reversal of the lower frequency, representing the double-sideband signal, will merely interchange the relative positions of the two peaks of either polarity. This change can have no effect upon the polarity or amplitude of the differential output.

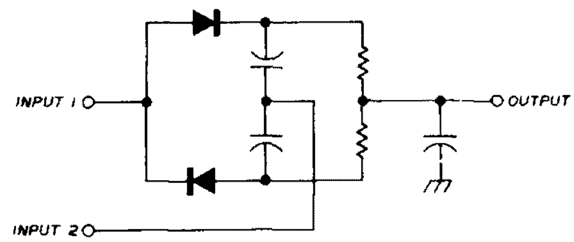


fig. 1. In this simple harmonic phase detector the two input signals must have a 2:1 frequency relationship. Applications include phase-locked loops and double-sideband signal reception.

A possibly over-simplified explanation would be to say that the 180° phase shift, inherent with the double-sideband signal, when compared with twice its frequency, is equivalent to a phase shift of 360°,

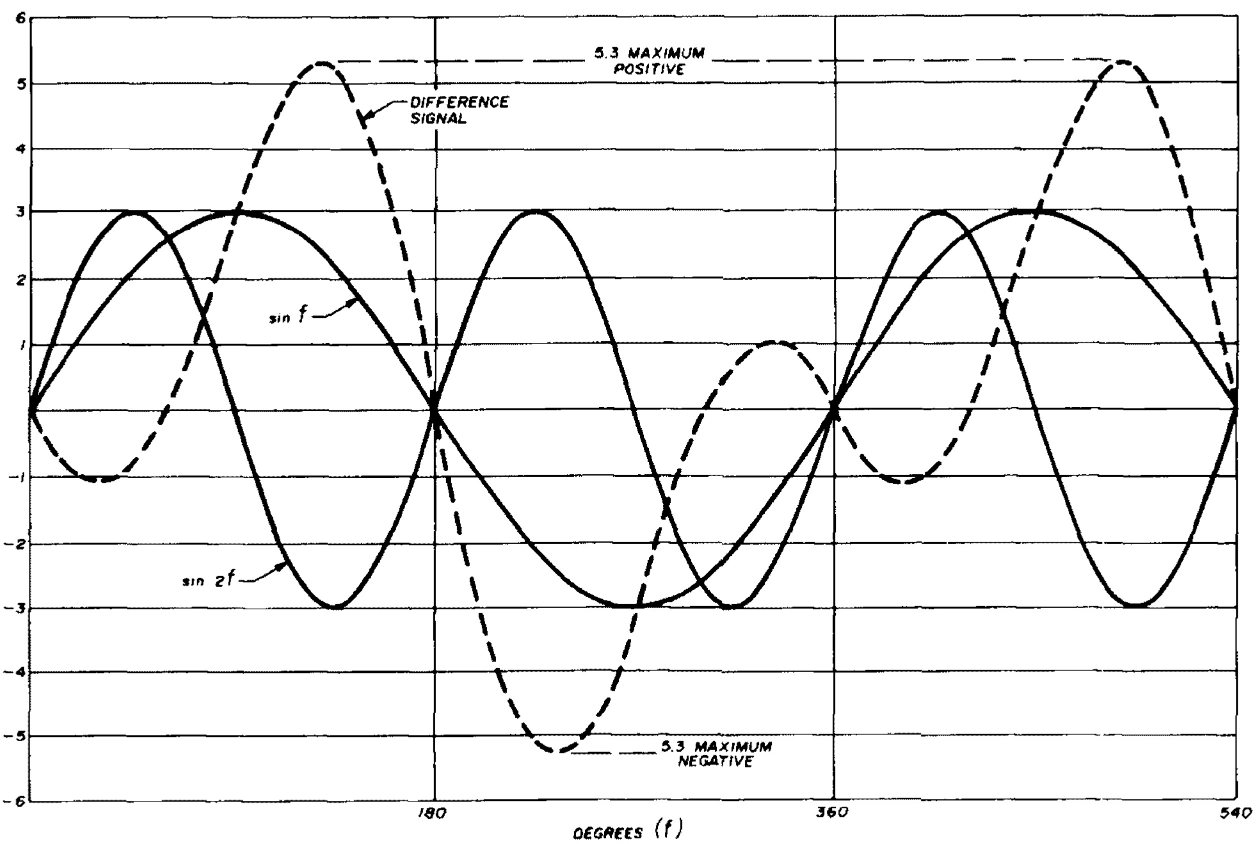


fig. 2. Operation of the harmonic phase detector with zero phase difference between equal signals at frequencies f and 2f. Differential output is zero.

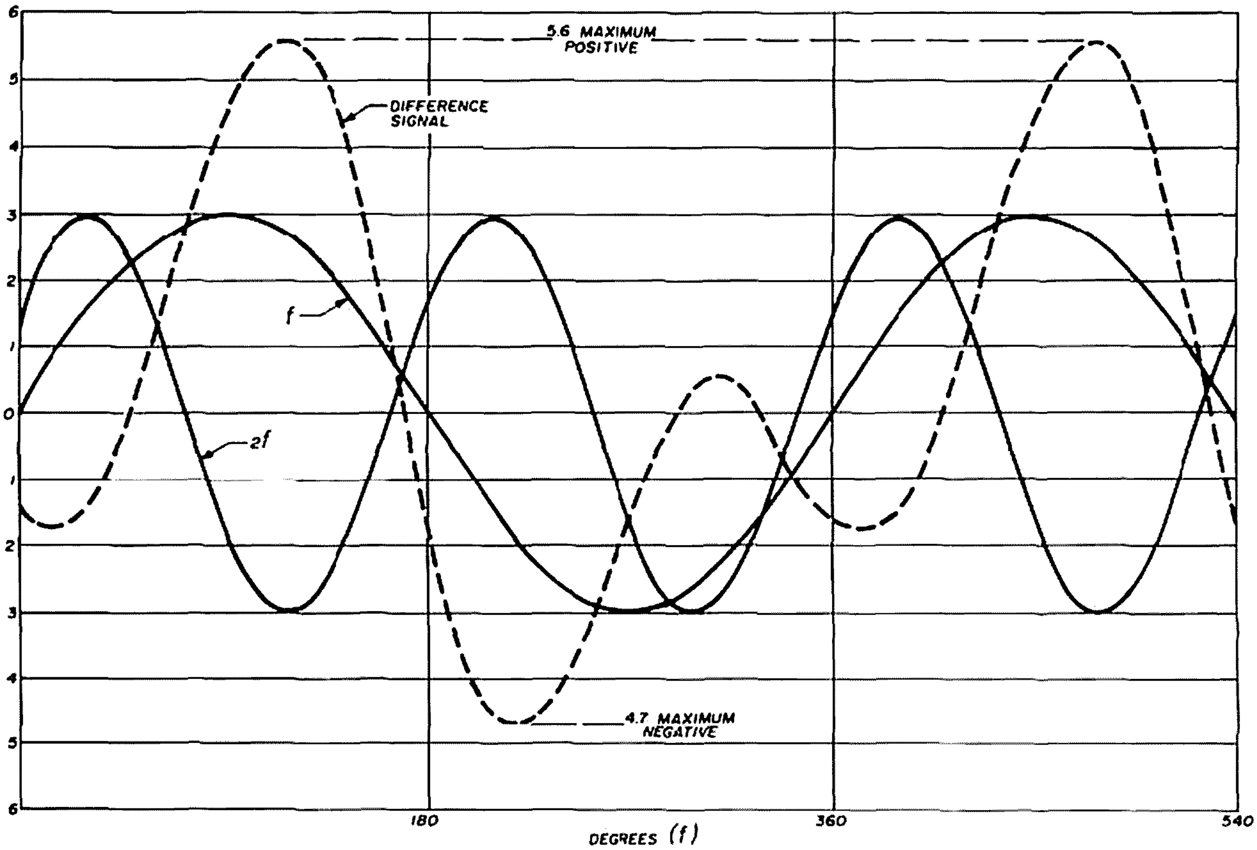


fig. 3. Operation of the harmonic phase detector showing effect of phase shift between signals at frequencies f and 2f. Differential output voltage is produced.

which is no phase shift at all! Suitable hard limiting of the double-sideband signal before its application to the phase detector, followed by suitable integration of the detector's output pulses permitted the necessary phase lock for double-sideband reception, but that's another story.

By grounding one of the two inputs (because the inputs are essentially in series) the detector may be made to function as an indicator of the most common form of second harmonic distortion of an audio-frequency signal, where limiting of either the positive or negative peaks is present. This fact suggests that the harmonic content of the human voice might conceivably be sufficient to lock the tuning of a ssb receiver with the aid of suitable filtering.

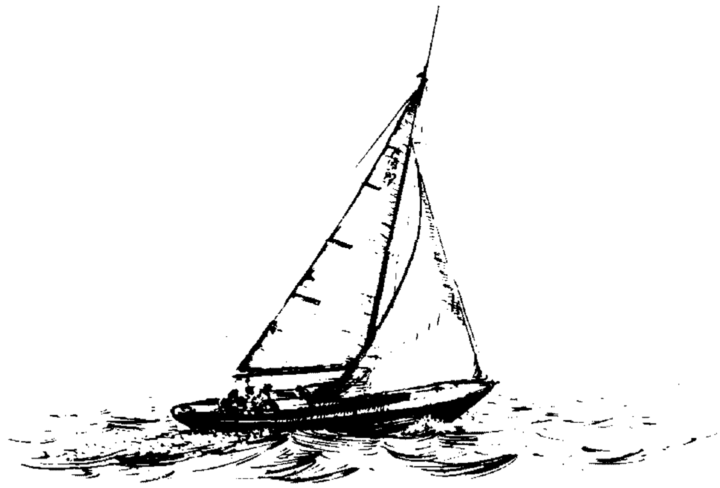
This simple circuit appears to have many possible applications which the amateur experimenter might find useful, and this article is prepared with that hope in mind.

"I wish you would learn to control your temper, and not try to punch everyone who says something you don't like."

reference

1. Stirling Olberg, W1SNN, "Reciprocating Detector," *ham radio*, March, 1972, page 32.

ham radio



amateur marine installations – small boat style

Overcoming the practical problems and legal restrictions of hamming from a pleasure boat

R.P. Haviland, W3MR, 136 Brownlie Road, King of Prussia, Pennsylvania 19406

Amateur radio station installations on private yachts of various sizes is increasing at a rapid rate. Operating a ham station on your boat is a nice way to combine your hobbies. However, it is also a hobby combination which has special requirements, both legal and practical. Let's discuss some of these special requirements, based on my personal experience as well as on observations of shipboard installations by other amateurs.

grounds

The radio ground system on a boat has three functions: rf grounding, lightning protection and corrosion prevention. For rf the rule is very simple—the larger the ground area, the lower the ground resistance and the greater the antenna efficiency. If it is large enough, the ground does not have to be in physical contact with the water—it can serve as one plate of a low-reactance series capacitor, with the water serving as the other plate.

The safest rule to follow for a lightning ground is to have lots of ground area in direct contact with the water, and a good direct-contact rf ground would also be a good ground for lightning. The lead wire should be large—at least number-10, and preferably larger. The radio and

lightning grounds can be kept separate, either by using a lightning arrestor or by having a heavy knife switch which grounds the antenna when it is not in use. One commonly used method is to make up a lead with a heavy clip on one end and a length of bare wire or a zinc electrode on the other, fastening the clip to the antenna and dropping the electrode end over the side during storms or when the boat is not in use.

There are two factors that must be considered in anti-corrosion protection. These are corrosion due to dissimilar metals, and corrosion from stray current flow. On most boats the regular anti-corrosion measures—zinc electrodes, plus cathodic protection on metal hulls—will also take care of any stray currents resulting from the radio installation. However, a good precaution is to replace the original power switches in the radio with a type which opens both sides of the supply, completely isolating the radio from the power source when it is turned off. It is also a very good idea to inspect all underwater metal on your boat a few weeks after making any new installation or change, to detect any problem before major damage occurs.

One very successful water contact ground is made from two copper tubes, typically 3/4- to 1-inch (1.9- to 2.5-cm) in diameter and 10-feet (3-meters) long, one mounted on each side of the keel. A variation of this uses a copper sheet, 3- to 4-inches (7.6- to 10-cm) wide by 10-feet (3-meters) long, tacked to the side of the keel or to the bottom of the boat. Both are good. On sailboats, the ground connection is often made to the bolts holding the lead keel. This seems to be satisfactory, but the ground lead should be connected to a number of the keel bolts to provide the lowest contact resistance.

On fiberglass boats a good non-contact or capacitance ground can be made by attaching 10 square feet (about one square meter) or more of light copper screen or perforated mesh to the hull below the waterline, using resin as the adhesive. This should also be satisfactory with wood boats, but I have never seen it so used. Some fiberglass sailboats carry the lead ballast inside the hull, as a casting set into the keel, and this can be used as a capacitance ground. All of these non-contact rf grounds should be supplemented by a lightning ground: I have seen an internal keel sailboat which was struck by lightning in which the charge escaped to the water by punching a number of small holes through the fiberglass at the upper edge of the keel casting.

The ground system of the sailboat on which I operate W3MR/M is a combination type. All shroud chainplates are connected by number-8 aluminum wire, a total of about 50 feet (15 meters). This makes a fairly good capacitance ground. In addition, two standard zinc teardrops on the outside of the hull are connected to this bus, further reducing the ground resistance, and giving lightning and corrosion protection. These zinc teardrops require replacement on occasion, showing that there is some current flow.

antennas

On many boats the mounting of an amateur antenna is complicated by the fact that the best antenna mounting place has been preempted by a marine radio installation. If a marine radio installation is already aboard, or one is contemplated, it should be remembered that the amateur station must be completely independent of and must not interfere with the marine installation. Compromises of the amateur antenna system, several trial installations, use of bandpass filters and the like, may be necessary.

Any of the mobile-whip or loaded-whip antennas can be used on a boat if one factor is considered. This is the extra lead length to ground, usually on the order of feet on a boat instead of inches as on an automobile. As a result, fixed-tuned loading coils will usually resonate outside the band. One solution is to use the loading coil for the next higher frequency band. Since the antenna is

actually being fed above ground, it has higher than normal impedance plus reactance. Even with adjustable loading coils, a matchbox is a good idea.

Some good mounting places for amateur-band whip antennas are: at the upper edge of the pilot house on the side away from the marine antenna; at the rail, usually on the stern quarter; at the masthead; or strapped to the mast, if it is wood. Try to keep a minimum of several feet of separation between the antenna and rigging, and as far apart as possible from other antennas.

Boats with masts can use wire antennas. On a power boat an antenna from the bow to a midships mast to the stern flagpole is good; feed can be at any convenient point, using a matchbox. On sailboats a standard installation is to insert large compression insulators at the top and bottom of the backstay, feeding it from the bottom end. Two-masted boats can have an antenna hung between the masts. My installation does this, using three separate lengths of wire cut for 10, 15 and 20 meters and operated as monopoles. The feedline coax shield is connected to the shrouds, which are grounded as noted above. In comparison tests with several other trial antennas this arrangement consistently gave the best results.

It is not really necessary to insulate rigging or wire to use it as an antenna. The objective is to get rf current flowing in a conductor that acts like an antenna. For example, on a typical medium-size sailboat the mast will be a 32- to 40-foot (about 10- to 12-meter) length of aluminum, a good approximation of a half-wave on 20 and a quarter-wave on 40. There are several ways to induce rf current flow. Some common ones are: feed the bottom end directly, if it is insulated (used on my boat on 40 and 80); use one of the shrouds as a gamma match; run a piece of insulated wire several feet (one meter) up the mast, again as a gamma match; or form the running-light leads (when present) into a coil, using this as part of a matching network. The key to success in the use of one of these methods is a good antenna tuner (see later)— and patience in making trials.

Beam antennas are rarely seen on small boats—they are too complicated and bulky, and not worth the trouble. If you want to try using a beam there are several short-element designs on the market, and a number have been described in various handbooks. Loading coils can be used to bring a TV antenna into resonance on 6 or 10 meters, although the bandwidth will be narrow. Another possibility is an array of three or four whip antennas, say on the wheelhouse or even on the quarters.

Still another possibility is a shore antenna. For temporary use a trap vertical lashed to a dock piling is good, though it may have to be retuned as the tide changes. It is often possible to use a long-wire antenna, end fed from an antenna tuner, and run as far as possible in the general direction of preferred signals. At the ship's home port a permanent rotary beam installation may be possible.

matching

If antennas other than trap verticals are used, it is a good bet that the antenna tuner will have to handle some unusual feed impedances. This eliminates use of some of the standard designs, which are intended for a limited range of loads and relatively low reactance.

After a number of trials I settled on a homebrew antenna tuner that is a modification of an ARRL Handbook design of some years ago. One or another of its switch settings will permit matching any impedance over the range from 3 to 30 MHz; the last position grounds the antenna. The tuner described in the May, 1974, issue of *ham radio* would be a good one for this application.[1] Because of the many possibilities, adjustment of the tuner can be tedious. It seems to be a good idea to try a pi configuration first. If this does not give a satisfactory match, try the low-impedance positions for loop-fed antennas, and high-impedance

settings for other types. Several configurations and settings of your tuner may give a match; use the one which is least critical to a change in frequency.

In addition to matching, an antenna tuner also helps keep the chassis of the transmitter at rf ground. Sometimes it will still be necessary to add filtering to the mike and key leads, and to set up special grounds for the transmitter chassis.

operating convenience

Because of space limitations, it is very difficult to get an operating position on a boat which is really convenient. The best I have seen was on a fair sized sailboat, where the place called "navigators area" on the ship's plans had been rebuilt into a radio area. Since this boat was used only in inland waters, the loss of navigational convenience could be accepted.

A suggestion—make a temporary installation, and use it long enough to find the type of operating you prefer—tied up or underway, phone or CW, net or casual, and so on. Then work out a convenient position for this type of operation. If you like net or favorite frequency operation, don't forget the possibilities of crystal control and a remote operating position.

legal matters

There seems to be considerable confusion about the proper identification when operating aboard a boat. The following are the tests I use:

If underway, operation is obviously *mobile*.

If at anchor or tied up, and the ship can get underway without interrupting a contact, operation is still *mobile*.

If some shore-side facilities are being used, and operation would have to be interrupted to get underway, operation is *portable* rather than mobile (typically when shore power or a shore-side antenna are used).

If the ship's location is in waters bearing the chart notation, "Use International Rules of the Road," the designator is still *mobile* followed by the ITU Region, rather than a location and/or call area. The Americas are in Region 2. Note that operation in coastal waters of another country requires permission of that country.

For the record, here is what the FCC says about your operation from shipboard:

97.101 *Mobile stations aboard ships or aircraft.*

In addition to complying with all other applicable rules, an amateur mobile station operated on board a ship or aircraft must comply with all of the following special conditions: (a) The installation and operation of the amateur mobile station shall be approved by the master of the ship or captain of the aircraft; (b) The amateur mobile station shall be separate from and independent of all other radio equipment, if any, installed on board the same ship or aircraft; (c) The electrical installation of the amateur mobile station shall be in accord with the rules applicable to ships or aircraft as promulgated by the appropriate government agency; (d) The operation of the amateur mobile station shall not interfere with the efficient operation of any other radio equipment installed on board the same ship or aircraft; and (e) The amateur mobile station and its associated equipment, either in itself or in its method of operation, shall not constitute a hazard to the safety of life or property.

Hamming from your boat can not only be a lot of fun, it can also enhance your boating safety by providing a much-needed communications link in time of trouble. Like any amateur installation in less than ideal circumstances, the typical shipboard ham station will be limited in both operating convenience and efficiency—but that won't keep any enthusiastic boater/ham off the air. Just listen for W3MR/M on all bands.

reference

1. Harry R. Hyder, W7IV, "Five-To-One Transmatch," *ham radio*, May, 1974, page 54.

ham radio

electronic speed control for RTTY machines

Single-knob RTTY speed control permits instant speed change

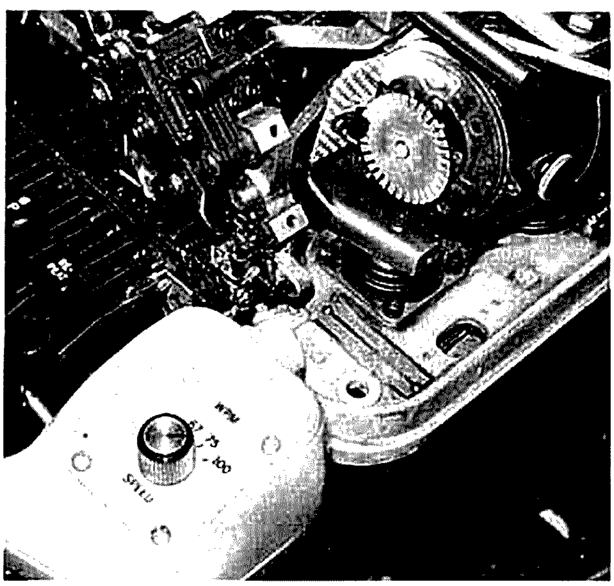
K.H. Sueker, W3VF, 110 Garlow Drive, Pittsburgh, Pennsylvania 15235

In a rash moment I decided to invest in a governor-motor Kleinschmidt TT271 KSR page printer to use for high speed operation. The prospect of 60 to 100 wpm operation for $40 was just too good to pass up. Much to my pleasant surprise, the governor worked perfectly and the machine generated no rf interference in my receiver. With 100 wpm gears and various governor adjustments it would run from 67 wpm to over 100 wpm and copy was fine at each standard speed. Then I decided to design and build a solid-state motor drive to replace the governor, permitting me to control speed from a console knob.

speed sensor

The first problem was to sense motor speed. For this I made a pulse wheel about 1-3/4 inch (4.5 cm) in diameter from 0.032-inch (0.8-mm) aluminum sheet (fig. 1) and, after removing the governor mechanism, fastened it to the motor shaft. The motor shaft had a hole which accepted a small machine-screw tap, allowing me to bolt the pulse wheel directly to the end of the shaft. A General Electric H13A1 optical coupler completed the motor-speed sensor. The coupler, which consists of a light-emitting diode (emitter) and phototransistor (detector) with a gap between them, mounted in a plastic housing, is designed for pulse-wheel operation. The emitter is excited continuously and the pulse wheel teeth interrupt the beam to the detector, as shown in fig. 1. I used a 33-tooth wheel with slots made by sawing 1/4-inch

RTTY speed control system mounted on a Kleinschmidt page printer showing one-knob speed control, front, and motor-mounted pulse tachometer, rear.



(6.4-mm) radial cuts with a coping saw. Fortunately, the wheel slots need not be made with precision.

circuit

The circuit is shown schematically in **fig. 2**. Pulses from the tachometer, squared up by transistor Q1, trigger a monostable multivibrator consisting of Q2 and Q3. The monostable multivibrator converts the tachometer output signal to a series of constant-amplitude, constant-width pulses with repetition rate proportional to motor speed. The average voltage resulting from this pulse train is also proportional to motor speed and so can serve as a highly accurate dc tachometer.

Operational amplifier U1 forms a three-pole Butterworth active filter which develops the required dc voltage from the pulse train. This filter configuration was chosen to give good ripple suppression with minimum time delay. Positive dc output current from U1, proportional to motor speed, is compared to a negative reference current derived from the speed switch and adjusting pots. Op-amp *U2* switches sharply from *on* (positive output) to *off* (negative output) when the tach-derived current exceeds the reference current. It serves as a highly sensitive speed detector and drives another optical coupler, this time an H15A1. This coupler is similar to the H13A1 but without the mechanical gap so sensitivity is much higher. The coupler switches transistor Q4 in the gate circuit of the triac which, at long last, turns the motor *on* and *off*. This second optical coupler isolates the control circuit from the 120-volt line.

Operation of the triac circuit is directly analogous to that of the governor. The motor voltage is either *on* or *off*, not continuously phase modulated. This permits considerable simplification in the gating circuitry and provides all the sensitivity needed for this type of application. The cycling rate is somewhat slower than that of the governor, however, probably due to the minimum *on* time limitations of a triac on a 60-Hz line. After working with the control circuitry for a while, you cannot help but admire the simple, rugged and sensitive mechanical governor with which these machines were originally equipped.

The resistors and pots in the speed reference network permit the circuit to

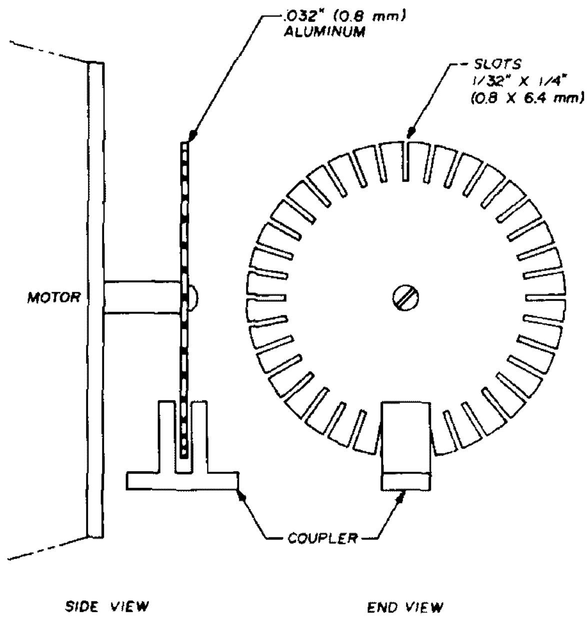


fig. 1. Plan view of the pulse tachometer. The author used a 33-slot wheel for his Kleinschmidt page printer, but suggests 60 slots if this controller is to be used with a Teletype Corporation machine.

be adjusted to each of the standard speeds of 60, 67, 75 and 100 wpm; these correspond to baud rates of 45.45, 50.00, 56.88 and 74.20.[1] Baud rates determine the precise speed ratios required, so a frequency counter can be used to set the speeds. In the Kleinschmidt machine, the motor shaft revolves at 3600 rpm for operation at the wpm speeds marked on the gears. Speeds may also be set by trial and error copy and that is how I did the job.

Operation over the range from 60 to 100 wpm is possible with the 100 wpm gears installed. However, the speed stability at 60 wpm leaves something to be desired. I ended up using the 67 wpm gearing and running the motor up to 5400 rpm for 100 wpm copy. This seems

to pose no particular problem for the motor since it will run much faster at full voltage, even under full load. I doubt that the additional bearing or brush wear is significant in amateur service. The 67 wpm gearing results in nearly normal shaft speed (roughly 3240 rpm) at 60

trigger pulses at the base of Q2. With the 15k feedback resistor in place, the monostable pulses at the collector of Q3 must not overlap at the maximum pulse-wheel speed. This may be checked with a scope or by observing the dc voltage at the collector of Q2.

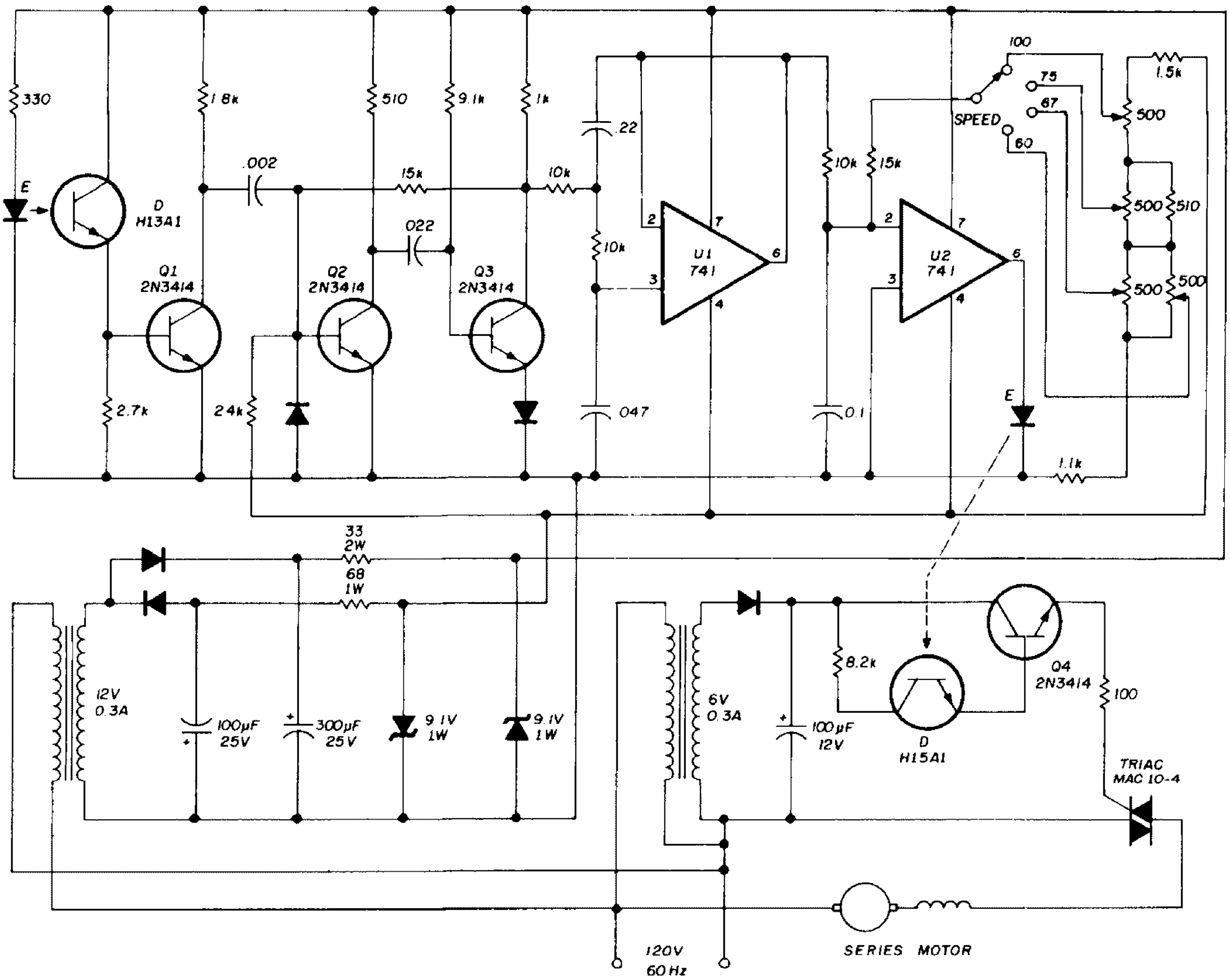


fig. 2. Schematic diagram of the RTTY speed-control circuit. Watch for built-in hash-suppression capacitors in the printer motor circuit, as they can destroy the triac motor driver.

wpm, and motor momentum serves to smooth out speed fluctuations caused by the rapidly clutched load.

adjustment

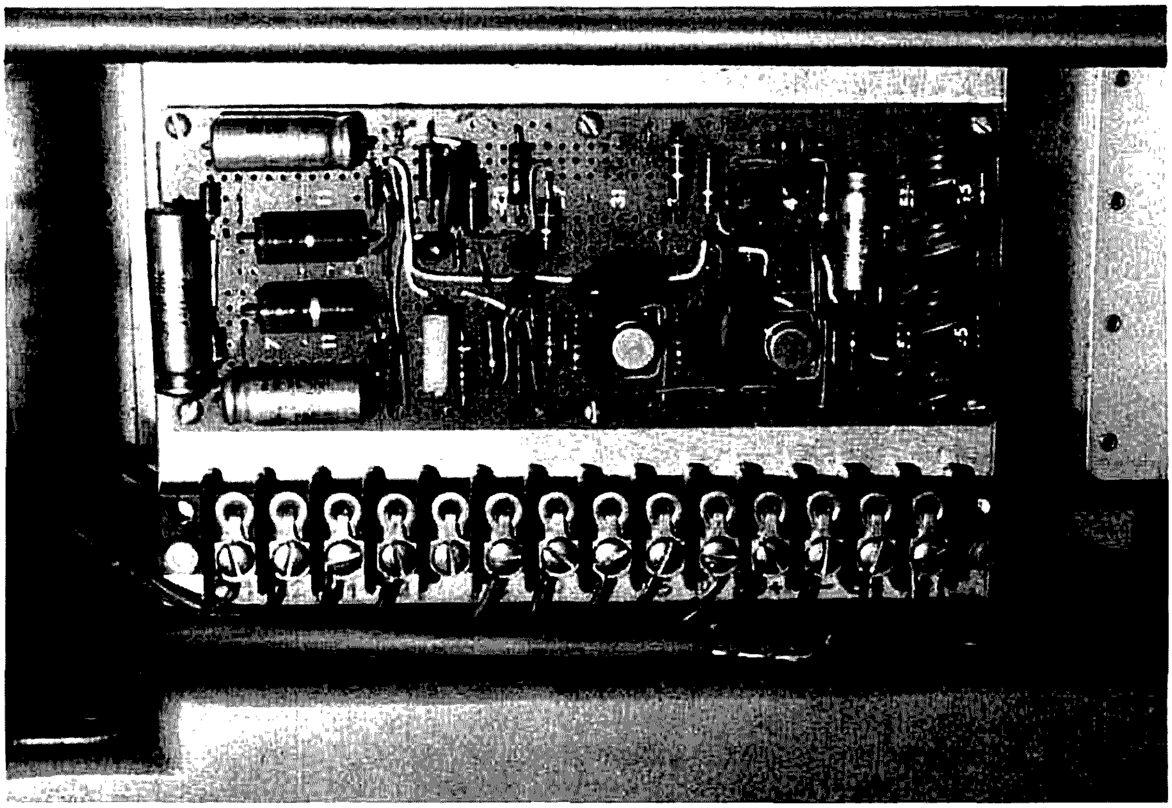
Some comments on adjustments are in order since no two pulse wheels or photo couplers will be identical. The 330-ohm LED dropping resistor and the 2.7k resistor in the base of Q1 may be changed, if necessary, to produce clean, steady

The voltage should be proportional to pulse-wheel speed over the desired speed range, reaching about 3 volts at maximum speed. Decreasing the 0.022-μF timing capacitor or decreasing the 9.1k timing resistor will narrow the pulses and reduce the average voltage. The best combination is the one which permits the maximum voltage while maintaining linearity to 100 wpm.

The range on the speed pots is deliber-

ately restricted to make adjustment easier. The values shown, together with the 1.1k resistor to ground, allow exact speed *ratios* to be set. Exact speed *settings,* however, may require changing the 1.5k upper divider resistor or the 15k op-amp input resistor to compensate for zener diode tolerances, resistor tolerances or gearing different from that described.

One precaution should be observed regarding the triac: Do not connect a capacitor, even a small one, directly across the triac or load. The surge current at turn-on can instantly destroy the triac. A resistor of at least 22 ohms should be inserted in series with any capacitor which may already be present or which you may wish to add for noise suppression. My unit caused some receiver hash while on the bench, but quieted down completely when installed in the machine and connected to the built-in rf filter. The MAC10-4 triac, mounted to the chassis with the insulating hardware supplied, should handle loads to at least 3 amperes. Higher current triacs or better heat sinking would permit the use of this speed control circuit with larger motors. Remember that the triac must supply 5 to 10 times normal current at turn-on while the motor is at zero speed. Also, a larger triac may require additional gate drive and a boost in the gate power supply.



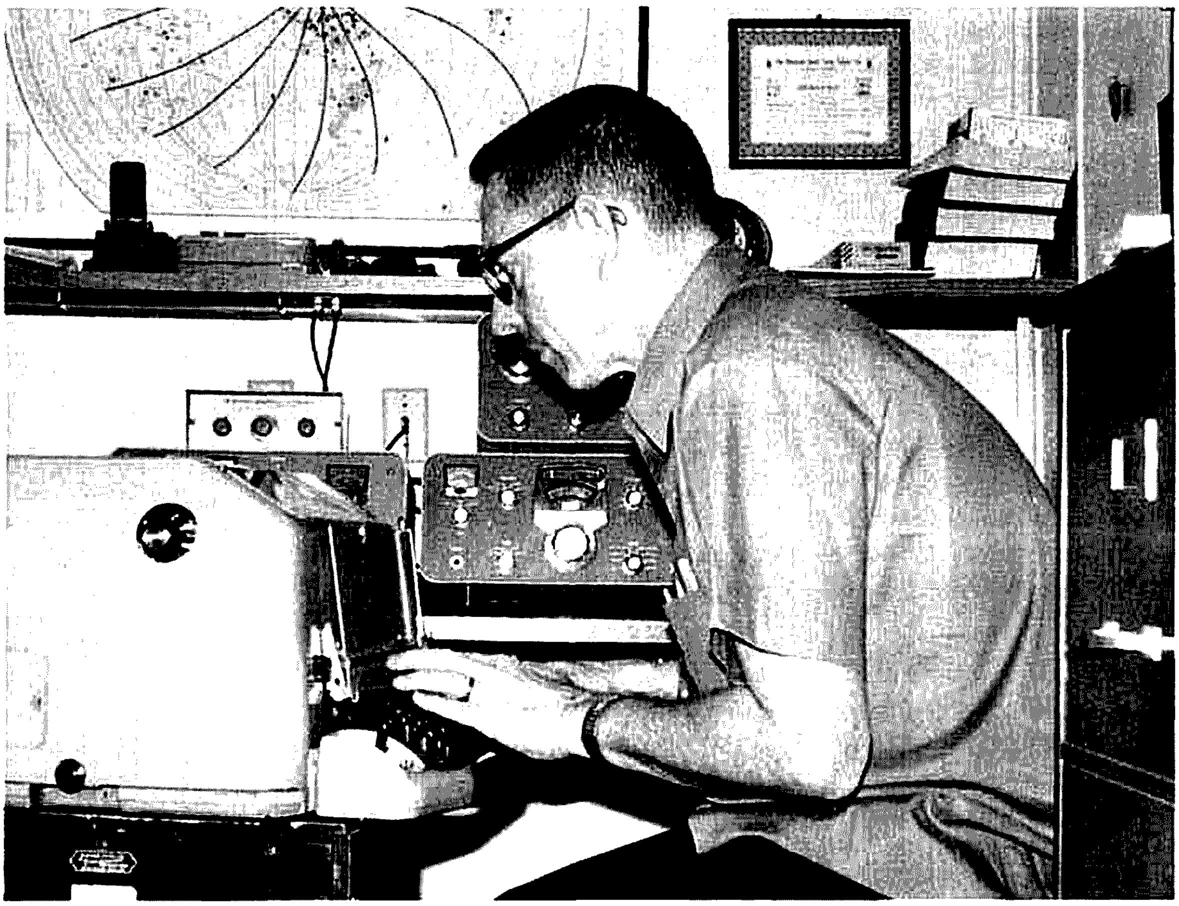
Circuitry for the RTTY speed control system is mounted on small perforated board which is installed on the rear of the page printer.

circuit components

None of the components used in this circuit are critical, though the timing resistor and capacitor should be types that are stable. Metal-film resistors and polycarbonate or polystyrene capacitors are suitable. If the GE photocoupler is not readily available, you can build your own using separate LEDs and phototransistors. Nearly any type of npn transistor is suitable for Q1 through Q4 (I used 2N3414s). The same is true for the diodes. Signal-type silicon diodes should be used in the transistor circuits, and lead-mount power diodes of at least

50-volt PIV in the power supplies. The op-amps may be individual 741 units or the 747 dual 741/741; pin numbers are shown for the TO5 type 741. Surplus units are available at attractive prices from various suppliers. The 6- and 12-volt power transformers are available from Radio Shack for about $1.50. Cost of the parts should not exceed $25, even if most are bought new.

I am not yet an expert on the Teletype Corporation machines except to note that the older ones, at least through the Model 26, have 1800 rpm synchronous motors. Teletype governor motors are in the same general speed range. This would suggest the use of a pulse wheel with at least 60 slots in order to get the tach frequency up near the design value. With that one change, the control unit described here should be suitable for Teletype machines with series governor type motors. It is *not* recommended for synchronous motors.


Author W3VF using the variable-speed Kleinschmidt RTTY machine at his station in Pennsylvania.

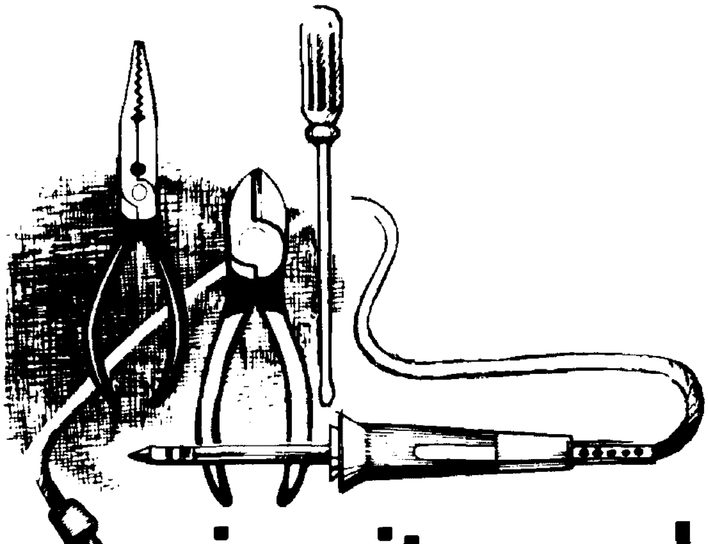
Replacement series motors are available from advertisers in *RTTY Journal* and *ham radio* for about $15.00.

In summary, my variable-speed Kleinschmidt machine has now been on the air for several months. Those patient hams who have coped with my typing have reported no speed problems, and the copy has been excellent on my end. Commercial stations have been copied at all four speeds with perfect results as far as speed is concerned. Reworking surplus machines and building controls of this sort may be viewed by some as an unrewarding chore, but the pleasure of achieving instant speed change by twisting a little knob made it all worthwhile for me. Now for that FRXD20 in the corner . . .

reference

1. "Principles of Telegraphy—Teletypewriter," NAVSHIPS 0967-255-0010.

ham radio



circuits and techniques

ed noll, W3FQJ

battery power

Amateur radio is entering a new era of battery operations. Battery-powered, hand-held and portable fm transceivers are just one example. Battery-powered QRP operation is also popular. Look for a new emphasis on battery operation during field-day activities. In fact, the very changeover from vacuum tube to all solid-state electronic gear has awakened new interest in battery applications.

The energy crisis is the catalyst for further expansion. The battery-powered bus has made a successful debut; the battery-powered family run-about waits behind the oil curtain. However, the real expansion is likely to be within the framework of solar power. Batteries will carry us through the night.

The all solar-powered ham station is no wild fantasy. The primary-cell battery is a one-shot affair. A typical example is the common zinc-carbon battery. After its chemical energy has been converted to electrical energy it is discarded. The secondary-cell battery is rechargeable. Its energy can be drained off and then resupplied by recharging the battery from a source of electrical power such as a solar-energy converter or a standard battery charger. The most well-known secondary battery is the lead-acid battery used in your automobile.

When your battery-powered ham gear is removed from your car, the two most common secondary batteries are the nickel-cadmium and dry gelatin-electrolyte, lead-acid types.

battery ratings

Be it a primary or secondary type, the two most common battery ratings are voltage and ampere-hours (or milliampere-hours). Voltage is no problem because there are a wide variety of types available according to voltage. There are many 12-volt types, matching the most common voltage requirement of solid-state radio gear. Other voltages can be obtained with the proper series-parallel groupings of standard-voltage batteries. The parallel connection, of course, increases current capability and available ampere-hours.

The ampere-hour or milliampere-hour ratings of batteries are usually based on 10 or 20 hours of continuous operation. For example, a popular 12-volt nickel-cadmium battery has a 1.2 ampere-hour rating. This is based on a 10-hour discharge time and a cut-off voltage of 11 volts. What continuous current demand could be made on the battery over this

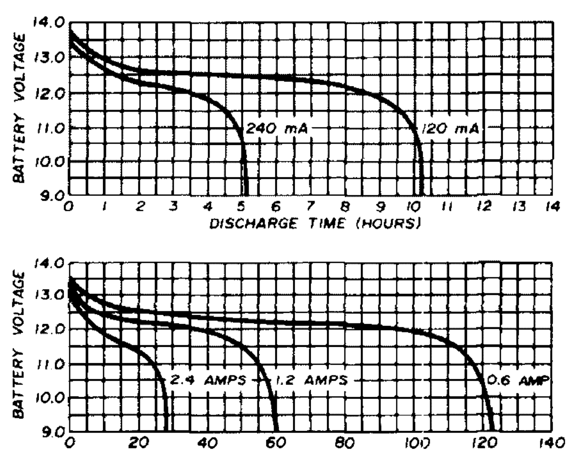


fig. 1. Discharge characteristics of a typical nickel-cadmium secondary battery (courtesy Eveready).

10-hour period? Since the ampere-hour rating of the battery and operating time are known,

$$I = \frac{\text{Ampere-hour rating}}{\text{Time}} = \frac{1.2}{10}$$

$$= 120 \text{ milliamperes}$$

The ampere-hour rating is less when there is a greater current demand and is likely to be more for a lesser current demand. Graphs and charts are available for various commercial batteries and from these you can determine primary battery life or when a secondary battery needs to be recharged. A typical graph is shown in **fig. 1**.

The upper curve shows the 120- and 240-milliampere discharge curves. Notice, on the 120-mA curve, that at the end of 10 hours the battery's voltage has declined to 11 volts. If the current demand is doubled to 240 mA, the 11-volt level is reached after something less than 5 hours, indicating that the ampere-hour capacity of the battery is somewhat less than its 1.2 rating for a 10-hour period. The bottom set of curves shows the discharge time in minutes when a high current demand is made on the battery. Note that, for a current demand of 1.2 amperes, the 11-volt level is reached after a time interval of 55 minutes.

An example of a battery chart is given in **table 1** for the popular D-size zinc-carbon (LeClanche) dry battery. Battery life is related to how many hours per day it is switched into operation. Take a current drain of 100 mA as an example. Note that, for 2-hours operation per day, the battery potential will drop to 1.2 volts after 29 hours. When the battery is operated continuously with a current drain of 100 mA its anticipated life to cutoff potential of 1.2 volts is only 9.6 hours.

table 1. Hours of life chart for Eveready 950 D-cell.

schedule	starting drains (mA)	cutoff voltage 0.8V	0.9V	1.0V	1.1V	1.2V
2 hours/day	10	525	500	475	450	430
	20	295	270	260	240	210
	30	210	185	175	155	135
	50	125	113	103	89	77
	100	57	50	45	35	29
	150	33	29	25	18.5	14
	200	21.5	18	15.5	11.5	8
	250	15	12	10	7.2	4.5
	300	11	8.5	7	5	2
4 hours/day	10	660	620	580	530	470
	20	330	310	290	260	230
	30	220	200	185	155	125
	50	123	108	96	81	64
	100	50	41	36	30	22
	200	18	13.5	12	9	5.2
	300	8	6	3.5	3	2
8 hours/day	10	700	660	620	560	460
	20	340	310	270	230	180
	30	210	180	150	130	100
	50	105	82	70	60	50
	100	39	28	23	18	13.5
24 hours/day	10	1050	745	600	500	370
	20	360	260	210	165	125
	30	200	145	115	88	65
	50	92	67	52	40	29
	100	32	24	18.5	13.5	9.6
	200	11.5	8.5	6.4	4.5	3.2
	300	6	4.5	3.5	3	2

The 1.5-volt LeClanche dry cells have been more or less standardized as to physical size and capacity. The data of **table 2** are typical. Ratings are based on an operating temperature of 70°F, two operating hours per day and a cut-off voltage of 1 volt for the first ten cells listed and 0.8 volt for the remaining cells except numbers 176 and 335, which are based on a cut-off of 1 volt.

table 2. Operating-hour capacities of standard Leclanche dry cells (courtesy Eveready).

cell	starting drain (mA)	service capacity (hours)	cell	starting drain (mA)	service capacity (hours)	cell	starting drain (mA)	service capacity (hours)
	1.5	275		0.4	210		2	190
N	7.5	52	105	2	30	143	10	40
	15	24		4	8		20	15
	2	290		0.5	435		2	510
AAA	10	45	108	2.5	103	145	10	105
	20	17		5	51		20	43
	3	350		0.6	710		2	560
AA	15	40	109	3	155	146	10	115
	30	15		6	75		20	50
	5	420		0.7	210		2	610
B	25	62	112	3.5	35	148	10	130
	50	25		7	12		20	70
	5	430		0.7	300		3	550
C	25	100	114	3.5	57	162	15	150
	50	40		7	25		30	65
	10	500		0.8	475		3	600
D	50	105	116	4	98	163	15	165
	100	45		8	49		30	72
	15	400		1	500		3	770
E	75	70	125	5	105	165	15	200
	150	30		10	45		30	90
	15	520		1	475		5	780
F	75	105	127	5	150	172	25	200
	150	45		10	72		50	90
	15	1000		1.3	275		5	1000
G	75	150	132	6.5	40	175	25	260
	150	65		13	16		50	110
	50	700		1.3	450		10	910
6	250	150	133	6.5	80	176	50	165
	500	70		13	35		100	63
	0.4	200		1.3	450		3	370
104	2	28	135	6.5	108	208-1	15	75
	4	7		13	52	·	30	35
	3	550		5	1000		10	1900
213-1	15	110	240-5	25	430	260-6	50	445
	30	50		50	180		100	210
	5	630		5	1800		10	375
240-2	25	270	250-5	25	750	335	50	78
	50	110		50	375		100	37

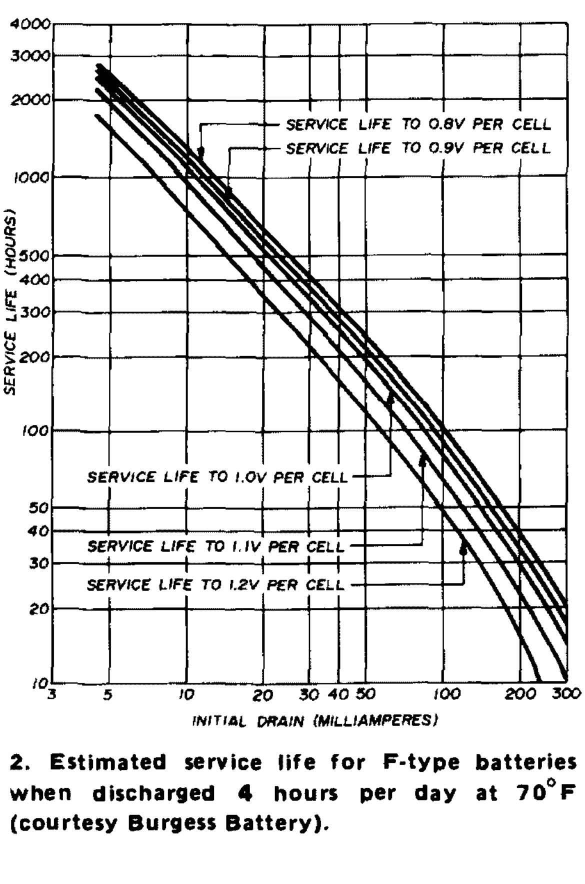
Use a D-size cell as an example. If a current drain of 100 mA for two hours per day is assumed, the service life will be 45 hours. This can be verified from the specific chart for a D-cell given in **table 1**. These basic cells are used to construct many of the higher voltage and higher current LeClanche dry-cell batteries. The F-type cell, for example, is popular in the construction of several popular communications batteries. To determine the service life when cells are connected

in parallel you need only divide the current drain by the number of parallel cells. Of course, the number of cells connected in series determines the final voltage and equals the number of cells times the voltage per cell.

A chart such as that shown in ***fig. 2*** gives more detail on the capacity of an individual cell type. In this example, service life in hours is plotted against current demand for various values of cut-off voltage.

What would be the service life in hours to a cut-off of 1.2 volts with a 50 mA current drain? From the 1.2-volt curve in **fig. 2** this can be determined as *120* hours. Figures are based on four hours of operation per day (this could be typical for many of the more active radio amateurs).

What would happen to the service life if two of these cells were connected in parallel, placing a current demand of only 25 mA on each cell? The 25 mA line in ***fig. 2*** intersects the 1.2-volt curve at approximately 270 hours. Note that this more than doubles the service life expected from a single cell.



2. Estimated service life for F-type batteries when discharged 4 hours per day at 70°F (courtesy Burgess Battery).

Temperature has a decided influence on battery discharge. In the D-type carbon-zinc cell example of **fig. 3** the decline to 1 volt occurs in about 1 hour at 70°F while the drop to 1-volt occurs at

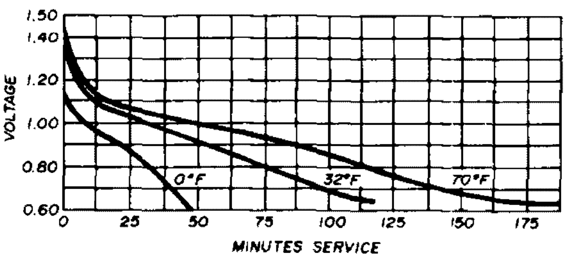


fig. 3. Temperature influence on rate of battery discharge (courtesy Eveready).

32°F in only one-half hour. Figures assume a current demand of 667 mA.

basic battery types

In experimenting with solid-state circuits and in QRPP operations of 1 watt and under, I usually use the common lantern battery. The 12-volt Eveready 732 and Burgess Radar-Lite TW2 consist of eight F-type dry cells in series. The data of fig. 2 are appropriate.

A current drain of 100 mA corresponds to an available power of 1.2 watts. If the lantern battery were subject to four hours of continuous operation per day the individual cell voltage would fall to 1.2 volts after 50 hours of operation.

This seems to be a short time. However, a number of additional factors must be considered. Do you operate four hours per day? How many hours per day would the 1.2-watt demand be made on the battery? This is important when you consider that your receive time is always longer than your transmit time. You do a substantial amount of listening around the band and then, depending upon whom you are in contact with, the transmit and receive times can be estimated at 50%. Furthermore, in CW operation you are placing an intermittent demand on the battery rather than asking for a continuous 1 watt of power. It may well be that during a 4-hour operating period your key may be *down* for one

hour or less. Thus, your actual battery life may be 4 to perhaps as high as 6 times greater than is suggested by the 50-hour life figure. This represents 200 to 300 hours of operating time and, in terms of the usual ham QRPP activity, you are talking about months of operation.

Two lantern batteries in parallel more than double the operating time. Or you can step up to the Eveready Hot-Shot 1463. This battery uses 16 G-type cells, table 2, connected in two parallel strings of eight. Since there are two sets in parallel, you must either halve the current values given under *starting drain* or double the *service capacity hours*, to obtain an approximate service capacity figure of the G-type battery data of table 2. This actual figure will be somewhat greater than the chart indicates because, as pointed out, the service capacity more than doubles when current drain is halved.

A rechargeable battery designed specifically for electrical and electronic applications is the alkaline-manganese dioxide type shown in fig. 4. Although rechargeable, their electrochemical systems can be hermetically sealed. Such batteries are

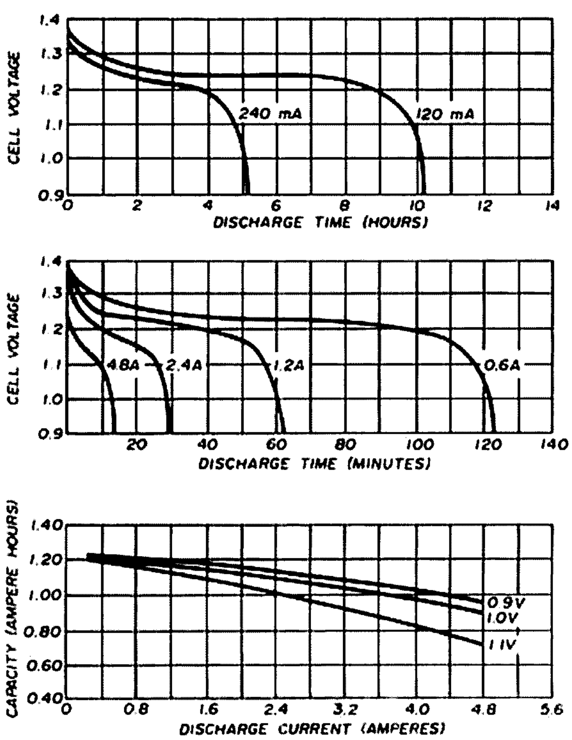


fig. 5. Characteristics of a nickel-cadmium cell (courtesy Eveready).

fig. 4. Rechargeable alkaline-manganese battery type 564 (courtesy Eveready).

maintenance free and they operate in any mounting position. Electrodes are made of zinc and manganese dioxide while the electrolyte is potassium hydroxide.

These cells come in two forms, meeting the physical dimensions of the D- and G-type dry cells. The characteristics of the Eveready D and G cells are given in table 3. Note the voltage values and the high ampere-hour capacities of these cells.

Higher voltage and/or higher discharge current capacities are obtained by using these fundamental cells in series and parallel groupings. This 564 battery, shown in fig. 4, uses nine G cells in series and has a 5 ampere-hour capacity that can deliver 1.25 amperes for a period of four hours without recharge. We are now talking about more than 10 watts of available power and, in fact, 10 watts continuous for a period of four hours. As mentioned previously, four hours of continuous key-down operation usually corresponds to more than 12 hours of continuous operating time.

Of course, you can anticipate many days of operation without a recharge. Considering that the battery can be recharged 25 or more times, there is adequate energy available to last a year or more for even the most active 5- to 10-watt QRP operator.

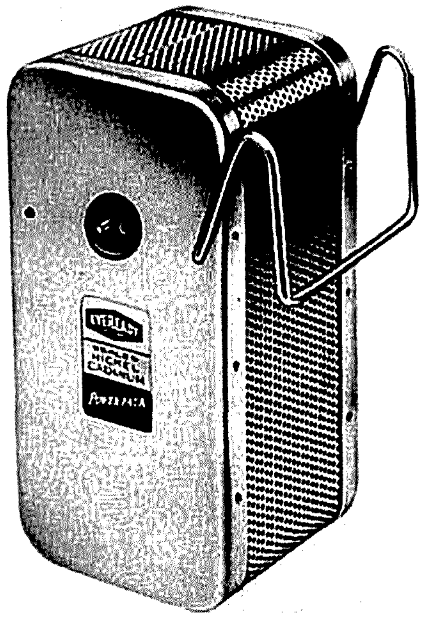


fig. 6. Sealed six-volt nickel-cadmium battery (courtesy Eveready).

The rechargeable nickel-cadmium battery is a popular secondary battery for use in electronics and electrical systems. Again, it is a shielded battery and there are no corrosive fumes or the need for adding electrolyte. An especially attractive advantage of the nickel-cadmium battery is its nearly constant discharge potential. The discharge-charge cycle may be repeated as many as 300 to 500 times.

table 3. Characteristics of basic alkaline-manganese cells.

cell size	nominal voltage	average operating voltage	rated ampere-hour capacity	maximum recommended discharge current
D	1.5	1.0 - 1.2	2.5	0.625 ampere
G	1.5	1.0 - 1.2	5.0	1.25 ampere

This means that a nickel-cadmium battery selected for most amateur applications will have an exceedingly long life it if is not abused.

Eveready nickel-cadmium 1.25-volt cells are available in button (20-300 milliampere-hours), cylindrical (150-4000 mA-hours), and rectangular (6-23 ampere-hours) types. Individual cells are welded together to obtain higher operating voltages. The cell voltage-discharge time characteristics of a typical cell are given in fig. 5. Note that the voltage on discharge holds rather constant over the discharge time until near the break in the curve. It is not advisable to let the cell voltage drop below 1.1 volt before recharge is initiated. In so doing the cell voltage is held up and the life of the battery is extended. The lowest graph shows the capacity-discharge current characteristic of the cell. For example, it shows that with a 1.2-ampere discharge current and a cutoff of 1.1 volt, the battery has an ampere-hour capacity of about 1.12 ampere-hours.

The Eveready N86 battery consists of ten of these cells connected in series to obtain 12.5 volts. On the basis of a 10-hour discharge at a rate of 120 mA, the capacity of the battery is 1.2 ampere-hours. Its discharge characteristic is shown in fig. 1. Note that the battery voltage drops to 11 volts near the 10-hour calibration line when the current demand is 120 mA.

There are many nickel-cadmium battery types. The Eveready 1007, fig. 6, is a four ampere-hour, six-volt battery that is completely encased. On a 10-hour basis it would have a rated current of 400 mA. It includes a socket for plugging in a companion charger. The Burgess CD33 is a 6-volter of lantern-battery size with a 2 ampere-hour capacity and includes a built-in charger.

The nickel-cadmium battery can be trickle-charged and kept up to full charge just as you would maintain an ordinary lead-acid storage battery. Usually the trickle-charge current is about one-quarter of the rated discharge current.

Next month's column covers battery chargers and the increasingly popular gelatin electrolyte lead-acid secondary battery. I will also discuss how a solar power converter can be used to trickle-charge or fully charge the various types of rechargeable batteries.

ham radio

the ham notebook

temporary fix for noisy volume controls

Gain control pots eventually become intermittent at the point of maximum use and thus can make a fine piece of equipment almost unusable. This happened to the audio gain control in my Collins 51J3, and having torn the receiver apart once before to work on the PTO I had no desire to go through all that work again just to change a noisy potentiometer.

Since a worn pot manifests itself by sudden jumps in level, distortion and bursts of noise, I reasoned that a "fix" that had a stabilizing effect at the tube grid might suppress the problem sufficiently to make the receiver usable again. Many audio circuits resemble the Collins circuit shown in **fig. 1**, so my first attempt was to put a fixed resistor from the wiper (tube grid) to ground. This did exactly what I wanted, and after a little experimentation to find optimum value (5k still provided enough audio to drive me from the room, yet reduced the effect of the noisy spot to the point where I have to listen carefully to find it) my faithful 51J3 was back in business.

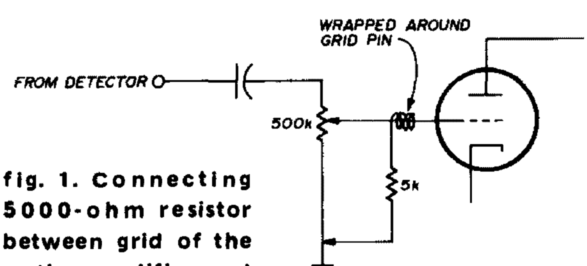


fig. 1. Connecting 5000-ohm resistor between grid of the audio amplifier and ground substantially reduced noise generated by worn 500k volume control.

Being essentially lazy, and seeing that this was a "temporary" repair anyway, I didn't even pull the receiver out of the cabinet to add the resistor. Instead, I simply wrapped a piece of wire around the grid pin of the audio tube, ran the wire up alongside the tube and out through the top of the shield, and soldered the 5k resistor to it. The other end of the resistor is grounded by a handy nearby wing nut. It may not look very nice, but it sure works well— and someday, when I *really* have to tear into the receiver, I *do* plan to replace that noisy pot!

Joe Schroeder, W9JUV

counted frequencies

In some configurations, a frequency counter may not properly show the received (or transmitted) frequency. I refer to those cases where a *clarifier* or variable bfo may be used. For example, in the Collins 75S-3 receiver the variable bfo changes the beat-note on CW. However, the 32S-3 exciter in transceive continues to use its own crystal bfo. Thus, a receiver frequency counter may not correctly show the transmitted frequency nor the frequency to which the receiver is tuned.

Similar peculiarities can occur in the use of the clarifier, a term used in Yaesu and some other designs. The nature of these devices should be studied before a counter frequency is relied upon to show the frequency that you are interested in.

Bill Conklin, K6KA

new products

two-meter fm sets

General Aviation Electronics (Genave) has introduced a new vhf fm system called *Mobiline*. This system, which consists of *Mobiline I* and *II* two-way fm mobile units and a variety of fixed base stations, offers considerable system flexibility. *Mobiline I*, for example, is designed for use in lightly populated areas with low signal density communications environments. These units accommodate two communications channels in the frequency range from 143.9 to 173.4 MHz, and are equipped with separate volume and squelch controls and transmit indicator light. A plug-in microphone and mobile mounting bracket with anti-theft device are included at no extra charge.

The *Mobiline II* provides the same basic capability as the *Mobiline I*, but has additional circuitry to permit use in heavily populated, high signal density communications environments. Refinements in the receiver section provide higher selectivity, improved spurious response and superior adjacent channel rejection. The *Mobiline II* like *Mobiline I*, features nominal 25-watts output power and 0.5 μV sensitivity for 20 dB quieting.

Either of these two radios is easily adaptable to fixed based station use. The Genave base-station packages include a stainless-steel 3-dB gain antenna with 50-feet of RG-8/U coaxial cable and connectors. Also included is the standard hand microphone (optional desk-type microphones are available starting at $34.95). Another important option is the *MobilPack* portability accessory. This allows the *Mobiline* owner to operate portable in the field or in vehicles without electrical systems. The *MobilPack* will provide a minimum of 24 hours of service on full charge and costs $124.95. Also available for the Genave *Mobiline* system is a sub-audible tone squelch system called *MobilGuard*.

The *Mobiline I* is priced at $319.95, the *Mobiline II* is $399.95 (including one channel). For more information write to General Aviation Electronics, 4141 Kingman Drive, Indianapolis, Indiana 46226, or use *check-off* on page 94.

test equipment for the radio amateur

Techniques of measurement and instrumentation have progressed rapidly in recent years, with the result that many very sophisticated methods of measurement are now available to the serious radio amateur. The development of modern semiconductor devices has meant that many items of equipment that were previously barred from the amateur workshop on the grounds of complexity, cost or size, have suddenly become a realistic possibility.

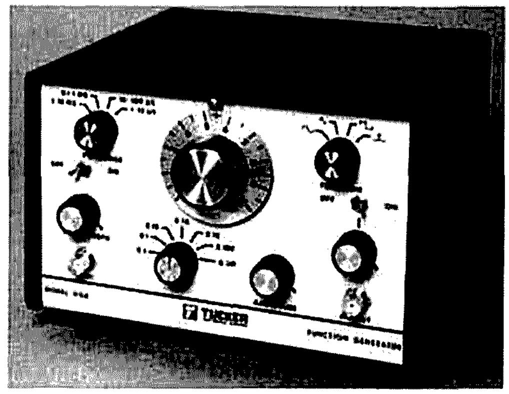
This new book by H.L. Gibson, G8CGA, published by the RSGB, explains in detail the theory behind modern measurement techniques, and goes on to describe many items of equipment of special interest to the amateur experimenter and home builder. Many of the construction articles feature instruments that have not before appeared in

print. Readers are taken through the basics of indicating and electronic instruments, and on to the techniques of frequency, power and noise measurement using up-to-the-minute components and methods. Included are chapters on meters, electronic instruments, dip oscillators, frequency and rf power measurements, noise measurement, antenna and transmission-line measurements, signal sources and attenuators, oscilloscopes, swept-frequency measurements, and components, vacuum tubes and transistors.

The section on electronic instruments includes circuit details on vacuum-tube voltmeters, fet voltmeters, transistor multimeters, diode probes and input attenuators. The chapter on dip oscillators includes valuable information on using the instrument as well as construction details for several solid-state units. The frequency measurement chapter covers frequency standards, crystal calibrators, digital counters, absorption wavemeters, harmonic indicators and a simple audio-frequency meter.

The chapter on rf power measurements includes data on dummy loads, thermal converters, rf ammeters, rf voltmeters, thermistor/bolometer bridges and directional wattmeters. Noise sources, noise generators, noise factor comparators and routine noise-figure checks are discussed in the chapter on noise, and vswr meters, vhf/uhf reflectometers, field strength meters and noise bridges are covered in the chapter on antenna measurements. Also described in the book are a rf signal generator, an audio signal generator, a two-tone test oscillator, oscilloscopes, modulation monitors, i-f sweep generators, an RCL bridge, a capacitance meter, a logic tester, a fet mutual conductance tester and many others. The reference data section includes much useful electronic and mechanical data as well as instructions for temperature and air-flow measurements. Hardbound, 132 pages. $5.95 from Ham Radio Books, Greenville, New Hampshire 03048.

function generator



Tucker Electronics Company has announced the first two instruments of what is expected to be a comprehensive new series. The model 300A is a 0.1- to 1-MHz basic function generator offering 10-volt p-p output into 50 ohms and less than 2% sinewave distortion. The instrument provides sinusoidal, square and triangular wave switched outputs and a corresponding sync output. Dc offset of ±2.5 volts is standard in the small, portable unit.

The second instrument is called the model 310A and adds variable pulse to the basic functions of the model 300A. The 310A's TTL compatible pulse can be varied from 1-μs to 10-millisecond pulse width and has rise and fall times better than 25 nanoseconds. All the standard 300A features are included.

In addition to an exceptionally low sale price, each instrument can be rented with an excellent purchase option. The Tucker 300A is priced at $195.00 (rental rate $19.50/month) while the Tucker 310A is priced at $295.00 (rental rate $29.50/month).

Tucker Electronics Company is best known as the world's largest distributor of reconditioned test instruments with an inventory exceeding 15,000 instruments. Tucker currently sells no less than 18 lines of new instruments including Weston Instruments, T.R.I. Corporation, Philips and several other well known lines.

For more information, write to Tucker Electronics Company, Post Office Box 1050, Garland, Texas 75040, or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

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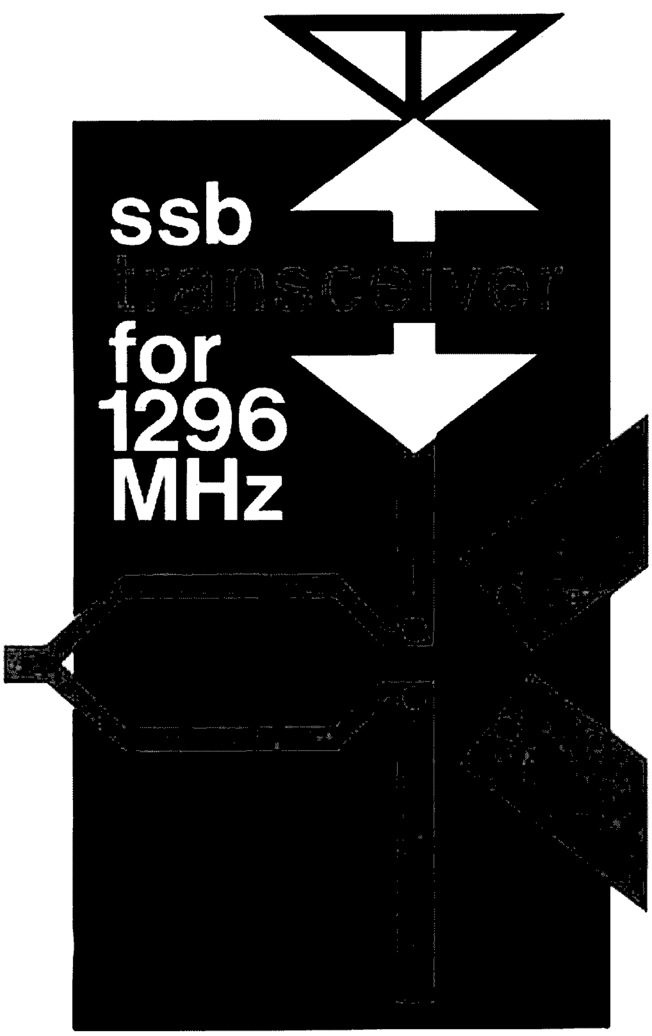
ham radio magazine

SEPTEMBER 1974



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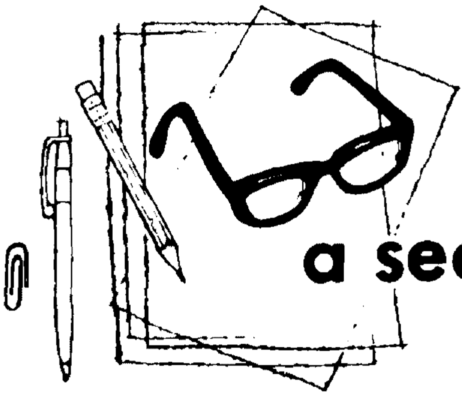


a second look

by jim fisk

As you look through this month's issue of *ham radio*, you'll see that two new sections have been added. The first, *stop press* on page 6, is a monthly summary of late news and happenings which affect amateur radio. In the jargon of the newspaper world, *stop press* means exactly that. If a hot story breaks while your morning paper is still coming off the presses, or if a big story takes a dramatic new turn, the presses grind to a halt so the new story can be added. If the story is big enough to justify a *stop press* edition, it's usually big enough to require a complete remake of the front page, banner headlines and all. Nor does the task stop there—room has to be found on the inside pages for stories crowded off the front page, re-edited to fit the available space, less important news items cut to the bone, and others discarded. Such a last-minute, crash effort is obviously expensive, so you don't see *stop press* editions every day, the editors opting instead for their regular early and late editions.

For a monthly magazine such as *ham radio*, however, a *stop press* edition in newspaper terms would be a practical impossibility. A crash effort to get WA6UAM's 1296-MHz article into this issue, for example, started the middle of May, immediately after it was first introduced at the West Coast VHF Conference. Nevertheless, it is possible, with proper planning, to insert whole pages a few hours before the magazine goes to press. Such is the case of *stop press* in this issue. With this technique, *stop press* is dated only by the slowness of second-class mail—typically three to four weeks. For immediate, up-to-date coverage of amateur radio news, in much more detail than can be presented within the limited confines of *stop press*, subscribe to *HR Report* (see page 98), our bi-weekly newsletter which more and more amateurs are depending upon to stay current with fast-breaking news from the FCC, ARRL and industry.

Some readers may be concerned that *stop press* is the beginning of an editorial swing to the non-technical aspects of amateur radio. Such is not the case—*ham radio* will continue to bring you the latest technical and construction information available, as we have in the past. But with a World Administrative Radio Conference scheduled for 1979, restructuring of the amateur service probably only a matter of time, and the many new proposals being offered by the FCC, it is imperative that the amateur community be well informed. *Stop press* is a small step in that direction.

The other new section in this issue is *pr bandstand* on page 76, a one-shot chance for our advertisers to blow their own horns. If you like the idea, we'll do it again next year. Our regular *new products* section, displaced temporarily by the *bandstand*, will return next month.

Jim Fisk, W1DTY
editor-in-chief

presstop

FCC ORDER RELAXES LOG KEEPING, and from now on NO mobile logs and only minimal fixed-station logs are required of amateur stations. Fixed (and portable) stations must maintain log book, but must record only contacts involving third-party traffic or during which an operator other than the station licensee was at the controls.

ANY EXTRA-CLASS LICENSEE COULD REQUEST PREFERRED CALL, even a two-letter call that previously required 25 years as a licensed amateur, under FCC Docket 20092. Extra Class licensees would also be permitted to hold as many "preferred" (one-by-three) calls as they had additional stations, although only one two-letter call would be permitted.

ARRL NATIONAL CONVENTION OPENED JULY 19, and for three days New York's Waldorf Astoria resembled the DX pileup on Kingman Reef. Weather was perfect, and show management estimates over 4000 attendees showed up.

Banquet Tickets Were Gone early Saturday afternoon, and commercial exhibitors were almost all enthused, as the three exhibit halls were well filled almost all the time they were open. Despite the high level of activity in the exhibit halls, however, the many forums all attracted substantial support too.

ARRL Band Plans for 50- and 450-MHz bands were announced at Repeater Forum. In brief, 6 meters will be low in, high out, on 20-kHz centers; 450 will be high in, low out, on 50-kHz centers. Auxiliary links, ATV and satellite communications are all accomodated on 450.

FCC CHAIRMAN RICHARD WILEY provided one of the weekend's high points with his ringing pro-amateur radio speech. Anyone who came with reservations about the new chairman's feelings toward the amateur service lost them quickly -- he's been looking us over pretty closely, and is, for the most part, enthused about what he's seen.

After Reviewing Recent Amateur-Related FCC Activities he touched on amateur restructuring (still evolving, and amateur inputs welcome) and Class-E CB (not yet decided one way or the other). He and the other FCC staffers then fielded questions from the floor: License fees -- will probably be reduced, perhaps to as low as half the present schedule. 160 meters -- should see some restrictions lifted soon in some areas as Loran A begins to be phased out. Codeless license -- very probable part of restructuring, but only above 144 MHz and placed so as to offer little disturbance to existing amateur operations.

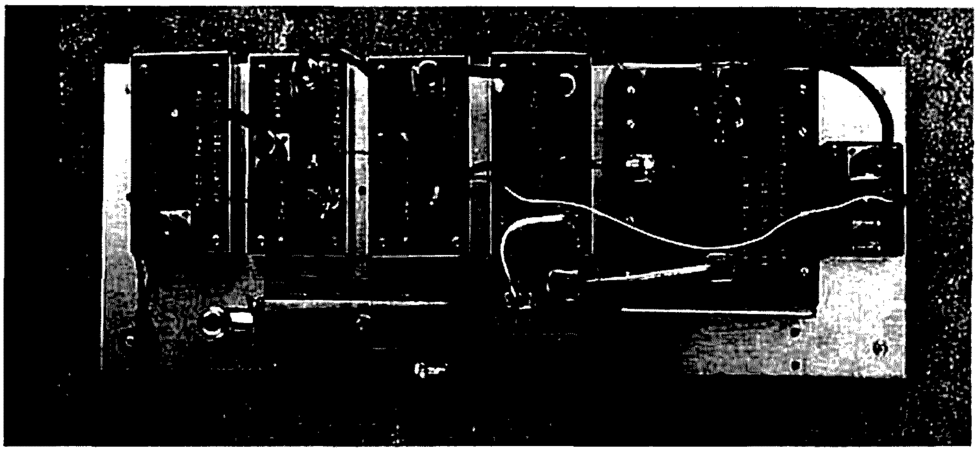
IMPORTANT NEW FCC ACTIONS highlighted the FCC Forum. FCC has just issued Notices of Proposed Rulemaking which would: lift the restrictions on repeater linking; lift restrictions on repeater crossbanding; and clarify commemorative callsign procedures.

NEAR RESTRUCTURING OF CB by FCC is provided by three items released July 24. Most far-reaching is Docket 20120, a Notice of Proposed Rule Making that greatly expands Class D (27-MHz) CB. 27.23 to 27.54 MHz segment would be added, with 26.96-27.31 to present AM and SSB only on 5-kHz channels! Restrictions on permitted communications would also be eased. This NPRM seems to neatly avoid or at least effectively put off action on the 220-MHz Class-E band. Hmmm...

The Second Shoe Dropped on CB July 24 is Docket 20118, which would prohibit the sale, lease, offer for sale or lease, or importation of RF power amplifiers capable of amplifying signals in the 20-40 MHz range! The only noteworthy exception to the ban is for multiband amplifiers specifically designed for amateur use.

Low-Power CB Handie-Talkies Will Move near 50 MHz if FCC Docket 20119 goes through. Commercial hand-held units not requiring operating licenses would be moved from 27 MHz to 49.91-49.99 MHz, and would require type acceptance by the FCC.

W5BWQ OF AUSTIN, TEXAS, is ham radio's most happy fellow, having recently taken delivery of a 1974 Chevy Vega, plus a car full of Swan amateur gear including an SS-200 Solid-State HF Transceiver and an FM-2XA Two-Meter FM Transceiver. All of this was his reward for winning HR's 1974 sweepstakes. Now it's your turn -- next year.



easy-to-build ssb transceiver for 1296 MHz

Complete description of a simple sideband system for the amateur 23-cm band — the same technique can be used for the other uhf bands

H. Paul Shuch, WA6UAM, 14908 Sandy Lane, San Jose, California 95124

The simple microwave ssb system presented here was used to achieve Northern California's first recent two-way ssb communications on 1296 MHz, between WA6UAM and K6UQH, on April 14, 1974.* Aside from any precedent which may have been established, the method used to transmit and receive microwave ssb represents a significant breakthrough in that it is simple, straightforward, inexpensive and readily reproducible by any uhf enthusiast. Neither specialized tools nor elaborate test equipment is required to build this equipment – equipment that provides the capability for line-of-sight ssb contacts on the amateur 23-cm band.

Fig. 1 shows the scheme generally used for the transmission and reception of uhf ssb signals. The received signals are down-converted in the conventional manner into the high-frequency spectrum where they are detected by the station receiver. Similarly, a high-frequency signal generated by the station exciter is heterodyned up to uhf, then amplified and transmitted. Note the high degree of redundancy present in this system. Both the transmit and receive converters use a mixer and local-oscillator chain, the function of each being essentially identical to its counterpart.

Assuming the availability of a bilatera

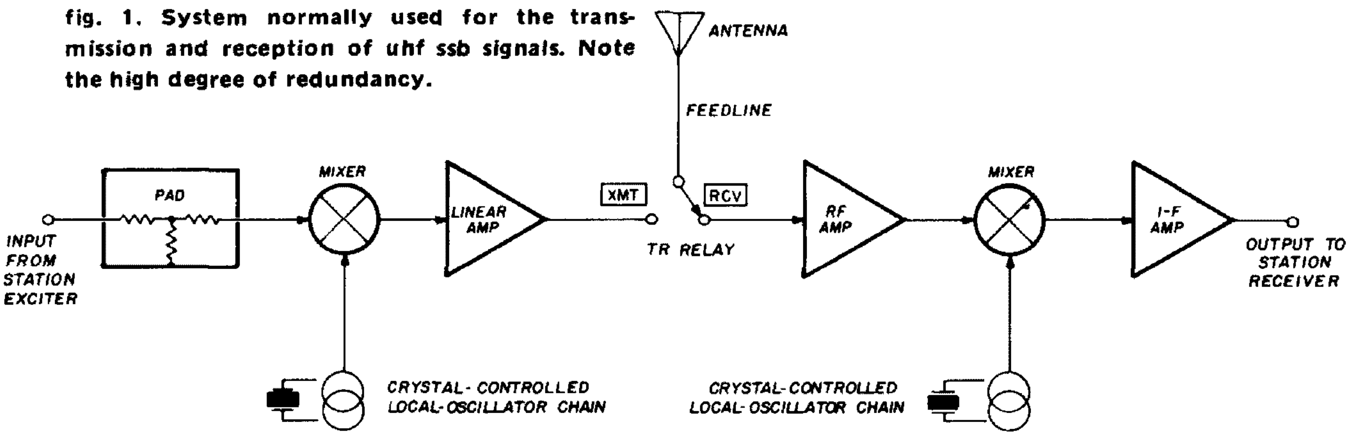
*Some years ago, K6HCP and W6GDO rar successful 1296-MHz ssb schedules, each using a 2C39 as a simultaneous mixer and LO doubler. The resulting ssb output was easily copyable, if not exactly spectrally pure. Bandpass filters were used to attenuate the undesired products at the antenna. The equipment currently used by WA6UAM and K6UQH, described in this article, uses diode balanced mixers with injection at the ultimate LO frequency. This method of heterodyning produces clean 1296-MHz ssb without excessive intermodulation products. Editor

mixer (one which operates equally well in both the *forward* and *backward* directions), the system can be simplified as indicated in fig. 2. Obviously, a passive mixer must be used in this application. Any active device designed to provide conversion gain in, say, the *up* direction, cannot function as a *down* mixer as well. Fortunately, singly- and doubly-balanced diode mixers function effectively in either direction, with only a few dB of conversion loss.

The greatest drawback of the diode mixer, so far as transmit conversion is concerned, is its limited power-handling capability. This normally requires the use of numerous stages of linear amplification *after* the transmit conversion.

It is evident from fig. 2 that by eliminating redundant circuitry, the TR switching complexity has increased threefold. Assuming that means could be found for eliminating the requirements for separate receive and transmit amplification, TR relays K1 and K2 could then be eliminated, too. Of course, such drastic simplification would jeopardize both the receive sensitivity and transmit power. However, depending upon the application, these tradeoffs might well be justified. Such was the case with the 1296-MHz station at WA6UAM.

fig. 1. System normally used for the transmission and reception of uhf ssb signals. Note the high degree of redundancy.



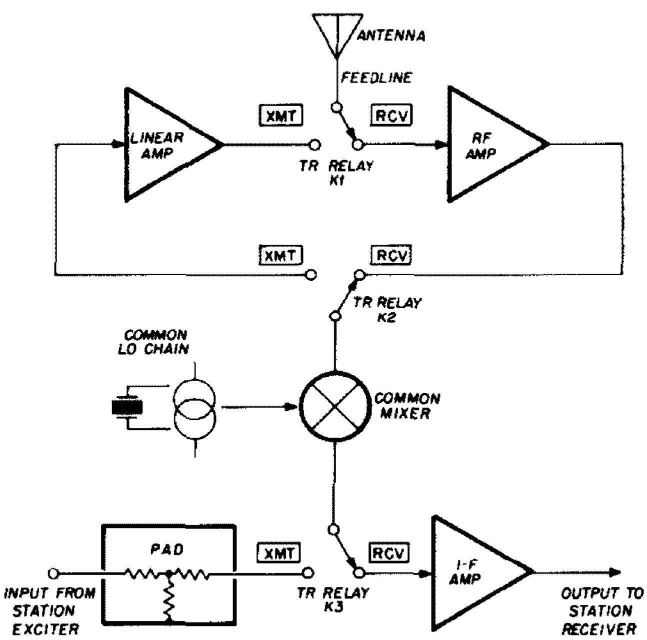


fig. 2. This uhf ssb system, which uses a common local-oscillator chain and balanced diode mixer, minimizes circuit redundancy but requires three TR relays.

The details for the Simple 1296-MHz Sideband System are shown in fig. 3. Equal emphasis was placed on simplifying the system to its minimum required content, and optimizing each subassembly to provide reliable communications over a reasonable range. Free-space loss and receiver noise calculations indicate that ssb communications between two such stations would be practical to distances of at least 100 miles (160 km).

Note the total elimination of TR relays and feedlines (and their resultant losses) in the microwave portion of this system. This is accomplished by mounting the mixer, filter and LO chain directly at the antenna (readily accomplished, as these modules are both lightweight and relatively small), and pumping only 28-MHz energy (plus 12 Vdc for the local oscillator) up and down the tower. The rf modules and antenna might even be

combined into a single physical unit, as shown in **fig. 4**.

mixer

So far as design tradeoffs are concerned, the mixer is, by far, the most critical component of the Simple Sideband System. To obtain a reasonable receive noise figure, low conversion loss is of paramount importance. At the same time usable transmit rf levels dictate high power-handling capabilities. As will be shown in a minute, these two criteria tend to be mutually exclusive. With readily available Schottky-barrier (hot-carrier) diodes in a balanced mixer, the system seems to optimize at about 6-dB conversion loss, with 3 mW of usable output power. Don't scoff at these seemingly restrictive figures. Calculations (see page 21) will show that this type of performance is more than satisfactory for communications to the edge of the visual horizon, and perhaps beyond.

Several reproducible uhf balanced mixers have been published recently.[1,2,3,4] The balanced mixer presented here is based upon a design developed by W6FZJ, and currently used by him on 2304 MHz. Versions for 1296 MHz have been built by both WA6UAM and WB6JNN, and provide a considerable improvement over the single-diode trough-line or interdigital designs frequently used by amateurs in uhf transmitting and receiving converters. An improved version of the W6FZJ mixer, which uses a commercially available balun to match the rf port of the mixer to the 50-ohm transmission line, will also be described.

Whatever mixer design is chosen, the diodes you select will determine its conversion loss and power-handling capability. One high-power, low-cost device is the Hewlett-Packard 5082-2817.*

*Hewlett-Packard 5082-2817 hot-carrier diodes are available in small quantities for $1.50 each from any Hewlett-Packard sales office. Matched pairs (5082-2818, $3.25), and matched quads (5082-2819, $7.00) are also available. If you can't find a Hewlett-Packard sales office in the *Yellow Pages*, write to Hewlett-Packard, 1101 Embarcadero Road, Palo Alto, California 94303.

These diodes have a burnout rating of 4.5-watts peak, or 1-watt CW, and are capable of conversion efficiencies of better than −5 dB.

practical mixer circuit

The complete 1296-MHz mixer shown in **fig. 5** uses hybrid construction (discrete components on etched microstripline), making duplication relatively straightforward for anyone with access to PC board fabrication facilities. Given sufficient time and patience, you can even "etch" your substrate with an Exacto knife and straightedge. At least four such manual efforts have been completed to date, and performance is equal in all respects to photochemically produced versions.

The equivalent circuit shown in **fig. 6** will help to clarify the operation of the mixer. Rf energy injected into the delta (Δ) port is transformed by the balun so that there is a 180° phase difference between the signals applied to the two diodes. The diodes are effectively in series and of like polarity so that the applied rf simultaneously biases both diodes on and then off, for alternate half-cycles.

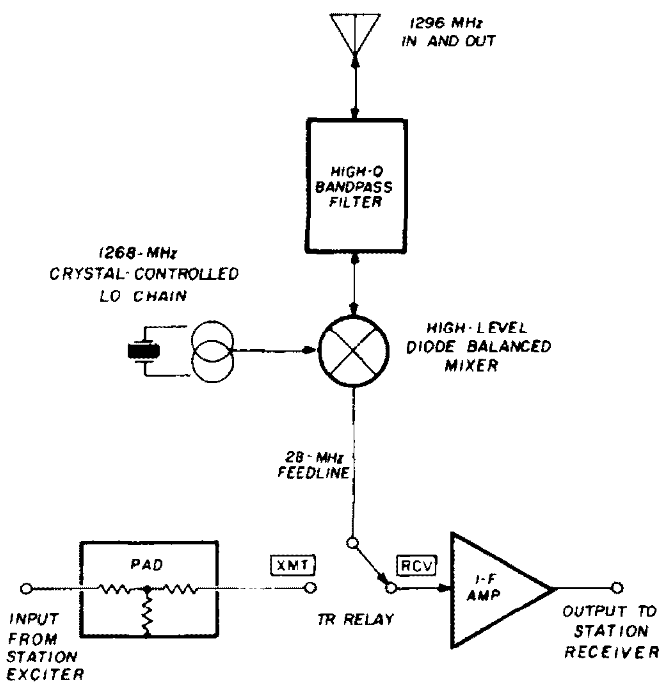


fig. 3. The Simple Sideband System, shown here, has been reduced to its ultimate simplicity but can still provide beyond-the-horizon communications when used with a high-gain antenna.

Rf energy applied to the sigma (Σ) input is transmitted down a quarter-wavelength, two-way power divider (which looks like a tuning fork) so that the signals it applies to the two diodes appear in phase. Since the diodes are in reverse polarity with respect to their bias return (common junction), this rf is applied to the two diodes out of phase. The simultaneous application of in- and out-of-phase rf signals to the diode pair results in a signal at their junction composed of the rf applied to the sigma (Σ) port, chopped at the rate of the rf applied to the delta (Δ) port. This complex repeating waveform can be shown by Fourier analysis to contain components of the sigma frequency, the delta frequency, their sum and their difference. Mixing, by the traditional definition, has occurred. The circuitry shown to the right of the diodes in **fig. 6** serves the purposes of signal conditioning (filtering out all but the difference-frequency component), dc bias return and a means of measuring diode bias current.

The most significant advantage of the balanced mixer over a single-ended design is that rf injected into the delta port is isolated from the sigma port, and vice-versa. To see how this is accomplished, consider a balanced signal applied to the two diodes through the delta port. In addition to feeding the diodes, this signal is shunted by the sigma port's power divider. Note that the power divider appears to this signal as a balanced transmission line shorted at the load end. Since this transmission line is a quarter-wavelength long at the delta frequency (assuming the delta and sigma signals are close in frequency), it transforms the short to an open, and the sigma port is effectively nonexistent so far as the delta signal is concerned.

Conversely, rf injected into the sigma port divides down the power divider, and appears to the diodes as two signals, equal in amplitude and phase. Looking toward delta, these two signals are cancelled in the balun and, thus, never reach the delta port.

It should be noted that single-balanced mixers provide no isolation whatever between the i-f port and either the delta or sigma port. Hence, filtering is required to remove the higher-frequency components from the i-f. Such filtering is accomplished in the hybrid balanced mixer by virtue of stubs at the i-f side of each diode, a quarter-wavelength long at 1296 MHz and open at the far end. These

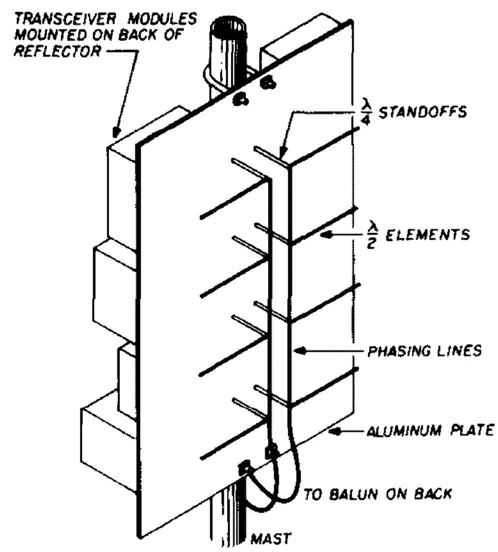
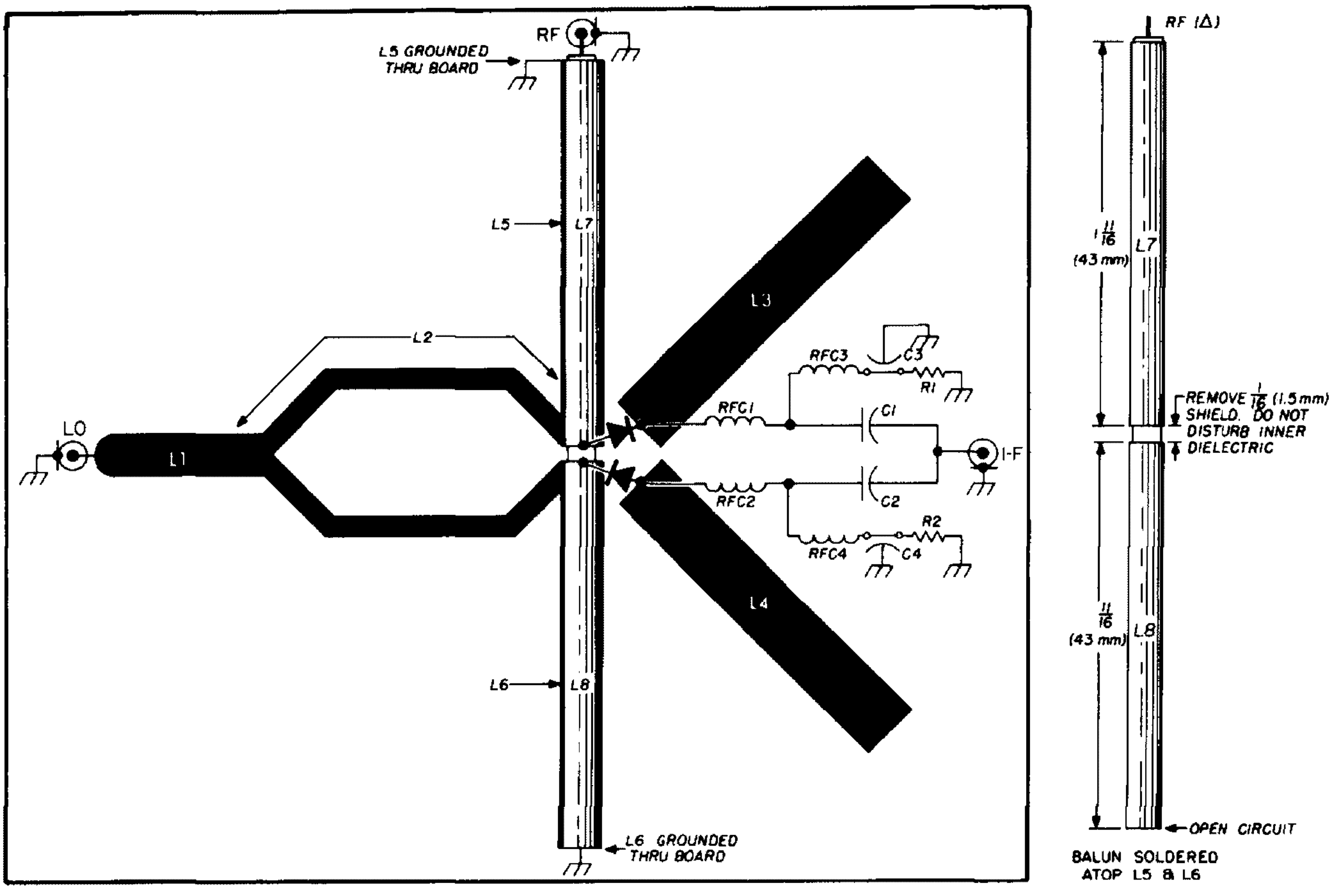


fig. 4. Antenna-mounted transceive converter. With this arrangement, using the system of fig. 3, only 28-MHz energy must be run down the tower to the operating room.

quarter-wavelength sections ground out the i-f port to energy near 1296 MHz.

Derivation of the balun used in the hybrid balanced mixer is shown in fig. 7. **Fig. 7A** is a coaxial balun frequently used in antenna work. In **fig. 7B**, the same balun is straightened out to improve symmetry. Note that a common ground is still required between the far end of the stub and a point one quarter-wavelength back on the feedline. This ground is frequently accomplished by connecting to the walls of the half-wave-long box in which the balun is built. In the case of the hybrid balanced mixer, the return is to a ground plane, on the opposite side of a substrate to which the balun is attached.

Both bandwidth and balance may be improved by modifying the in-line balun



C1,C2	0.01 μF disc ceramic
C3,C4	1000-pF feedthrough
CR1,CR2	hot-carrier diodes (H-P 5082-2818)
L1	50-ohm micro-stripline, 0.168" (4.5 mm) wide, any length
L2	75-ohm micro-stripline, 0.080" (2.0 mm) wide, 1-7/16" (36.5 mm) long, (along center)
L3,L4	38-ohm micro-stripline, 0.25" (6.5 mm) wide, 1-7/16" (36.5 mm) long
L5,L6	50-ohm micro-stripline, 0.168" (4.5 mm) wide, 1-19/32" (40.5 mm) long
L7,L8	50-ohm UT-141 coaxitube, 0.141" (3.5 mm) diameter, 1-11/16" (43 mm) long
R1,R2	10-ohm, ¼-watt carbon composition resistors
RFC1,RFC2	2" (51 mm) no. 32 wire, close wound on 0.050" (1.5 mm) diameter form or ferrite beads on leads of C1 and C2
RFC3,RFC4	22μH

fig. 5. This high-performance 1296-MHz balanced mixer uses etched 1/16" (1.5 mm) thick Teflon-fiberglass printed-circuit board and a coaxial balun. Full-size printed-circuit layout is shown in fig. 8. An equivalent circuit of this mixer, illustrating circuit operation, is given in fig. 6.

of **fig. 7B** as shown in **fig. 7C**. In this version the stub is a piece of coax identical to the original feedline. The center connector of the feeder is now connected to the center conductor, rather than to the shield of the stub. At the far end of the stub the center conductor is open. A quarter-wavelength toward the source (at the junction of the feeder and the stub) this open is transformed to a short, and rf sees the center conductor of the stub as being continuous with the shield. Therefore, the circuit at **fig. 7C** is electrically identical to that of **fig. 7B**, but with improved physical symmetry. Balanced output is taken from the same point as before.

Note that at the far end of the stub the center conductor must be open, and the shield grounded. Again the balun may be constructed upon a substrate, with return through it to a groundplane.

The balun used in the 1296-MHz mixer is made from a single piece of UT-141 type semirigid coax (50-ohms, Teflon dielectric, 0.141-inch [3.5-mm]

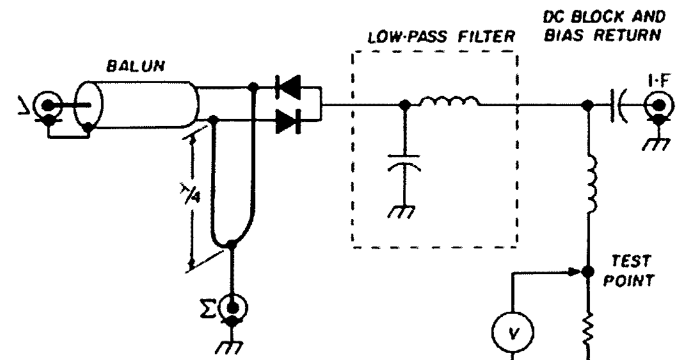


fig. 6. Equivalent circuit of the balanced mixer shown in fig. 5.

OD). Correcting for velocity factor, the quarter-wavelength sections are each 1-11/16-inches (43-mm) long. Judicious use of an Exacto knife and small tubing cutter will aid in the removal of 1/16-inch (1.5-mm) of the outer conductor at the junction of the feeder and the stub. Since no physical connection must be made to the center conductor at this junction, the Teflon dielectric should not be disturbed. Allow a short length of center conductor to extend beyond the quarter-wavelength section comprising the feeder coax. This will be attached to the center pin of the delta port's coax connector.

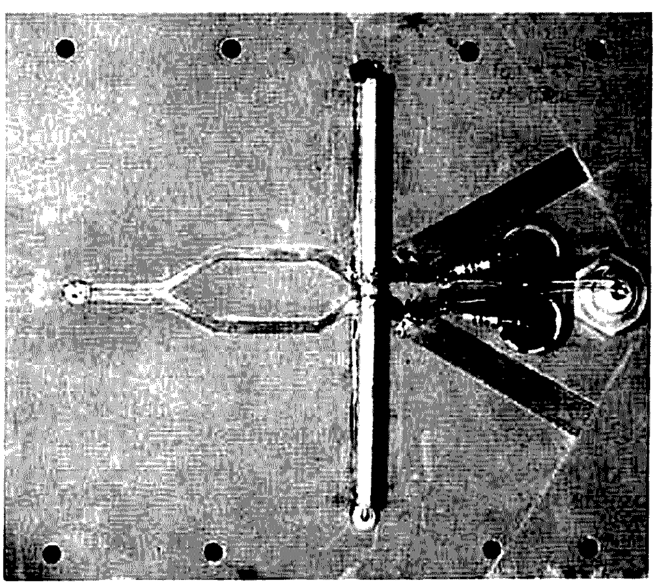
The mixer's substrate is etched on one side of a 1/16-inch (1.5-mm) thick, double-sided, 1 ounce, copper-clad, Teflon-fiberglass PC board. *Do not* use fiberglass-epoxy board, as its dielectric constant is not correct for the dimensions provided in fig. 5. The use of Teflon PC board is necessary in this design so that the velocity of propagation (and hence the electrical wavelength) of the striplines will approximate that of the coax balun. The full-sized PC layout is shown in fig. 8. All micro-striplines must be opposite a groundplane (the other unetched side of the double-sided board).

Although Teflon PC board makes an excellent substrate for micro-striplines at 1.3 GHz, it is quite expensive (and in some areas, totally unobtainable). The use of fiberglass-epoxy board, though it would increase losses slightly, would bring this type of equipment within reach of many experimenters who might otherwise be deterred. Of course, a glass-epoxy substrate is incompatible with the UT-141 coax balun because of the widely different velocities of propagation of the two mediums. In order to develop a 1296-MHz balanced mixer on glass-epoxy board, a different method of unbalanced-to balanced transformation is required.

Anzac Electronics manufactures an appropriate balun of moderate cost, excellent electrical performance and small physical size which frees the mixer design from restrictions as to substrate material.* Mixers built on 1/16-inch (1.5-mm) G-10 double-clad PC board using the Anzac balun exhibited improved matching at the rf port, an effect which more than offsets any additional losses in the glass-epoxy dielectric.

A schematic of the improved 1.3-GHz balanced mixer using the commercial balun is shown in fig. 9. Note that pins 1, 3 and 5 of the balun must be grounded *through* the substrate to the groundplane. When mounting the balun, do not allow its case to short out the striplines of L2.

*Anzac model TP-101, 500 kHz to 1.5 GHz, 50-ohms balanced to 50-ohms unbalanced transformer with midband insertion loss of 0.4 dB maximum and vswr 1.6:1 maximum, $15.50 in single quantities from Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154.



Construction of the doubly-balanced mixer of fig. 5, showing installation of the coaxial balun. Capacitors C1 and C2 can be seen between the two feedthrough capacitors.

Full-size artwork for a mixer board on 1/16-inch (1.5-mm) G-10 PC board (ϵ = 4.8) is provided in **fig. 10**. Either of these two balanced-mixer designs (**fig. 5** or **fig. 9**) will provide satisfactory performance in the Simple Sideband System for 1296 MHz.

output power - noise figure tradeoff

To avoid excessive intermodulation distortion in the transmit mode, it is desirable to inject into the i-f port a signal level at or below the mixer's 1-dB compression point. This is the level beyond which incremental increases in input power result in an ever-diminishing increase in output power. Such a situation typically occurs with i-f injection 5-dB below the local-oscillator signal level.

Due to the ready availability of any desired signal level at 28 MHz, the i-f injection level is not considered a limiting factor in system design. The following discussion assumes the use of the optimum i-f injection level in the transmit mode; that is, 5-dB below whatever LO insertion is applied.

In the transmit mode the usable output power is equal to the i-f injection level minus the mixer's conversion loss, L_c. For operation at the 1-dB compression point, this relationship can be expressed as

$$P_{out} = P_{LO\,(dBm)} - 5\text{ dB} - L_c$$

This formula implies that the system power output continues to improve for increases in LO injection. This would be true were it not for the fact that conversion loss does not remain constant for all levels of LO injection. **Fig. 11** demonstrates the variation in conversion loss, as well as optimum power output, as a function of LO injection level for a typical microwave diode balanced mixer. Note that optimum conversion loss occurs at an LO injection level of around +8 dBm.

Beyond this point, though conversion efficiency degrades, output power continues to increase. Indeed, within the power restriction of the HP 5082-2817 diodes, it is possible to obtain about 16 mW of clean output power. However, recall that the Simple Sideband System uses the same mixer and local-oscillator chain for both transmit and receive. Any

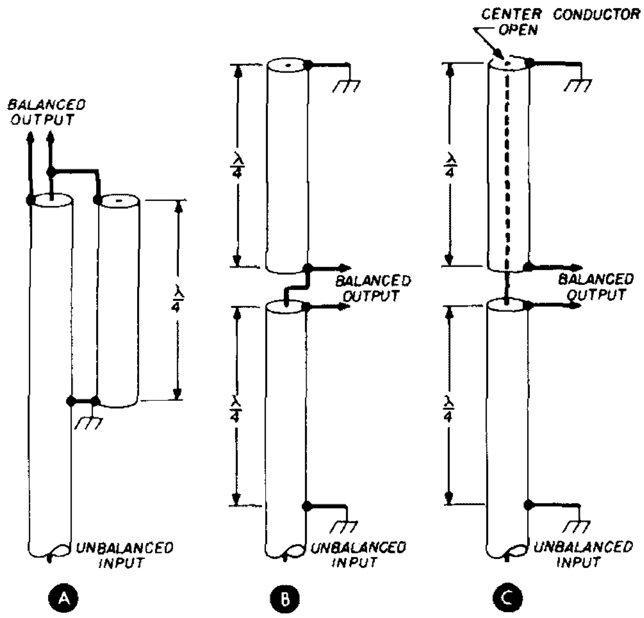


fig. 7. Derivation of the coaxial balun used in the balanced mixer of fig. 5. Conventional balun in (A) has been straightened out in (B) to improve symmetry. Balun in (C) is identical, electrically, to the circuit in (B) but symmetry has been further improved.

decrease in mixer conversion efficiency will degrade receive noise figure accordingly. Beyond +10 dBm of LO injection, transmitter power is gained only at the expense of receiver noise figure.

The break-even point occurs at an LO power of +16 dBm. Beyond this level, each dB of increase in transmitter output results in one dB of receiver degradation. Thus, the Simple Sideband System optimizes at 6-dB conversion loss, 40-mW of LO injection, 12.6-mW of i-f drive and 3-mW of output power. The tradeoff involved in determining this optimum performance point is illustrated by the $P_{out} + L_c$ curve of **fig. 11**. The sum of conversion loss and output power is used as a figure of merit for communications between two identical systems. This sum

represents the output power available at the i-f port of a receive mixer which is driven by an identical transmit mixer operating at the 1-dB compression point. Note that as local-oscillator power is increased, a knee is reached beyond which additional power will serve no useful purpose. Thus it is desirable to operate the system at the knee of this curve, which I call the *transceive figure of merit* curve.

bandpass filter

With 40 mW of LO injection in the transmit mode, and using the balanced mixer described above, a mixer output power of 6 mW is indicated on a power meter. It must be remembered that this signal represents both the desired output signal (LO + i-f) and the image (LO − i-f). A bandpass filter with sufficient skirt selectivity to reject the image will also have about 0.5-dB insertion loss in the passband. The actual transmitter output power, after the filter, will thus be

$$+7.8\,dBm - 3.0\,dB - 0.5\,dB = 4.3\,dBm = 2.7\,mW$$

Note that the filter will also eliminate receiver image noise, as well as blocking those out-of-band signals which might otherwise enter the mixer and cause cross-modulation distortion and interference. Of course, filter insertion loss must be added to mixer conversion loss and i-f noise figure when determining receive converter performance.

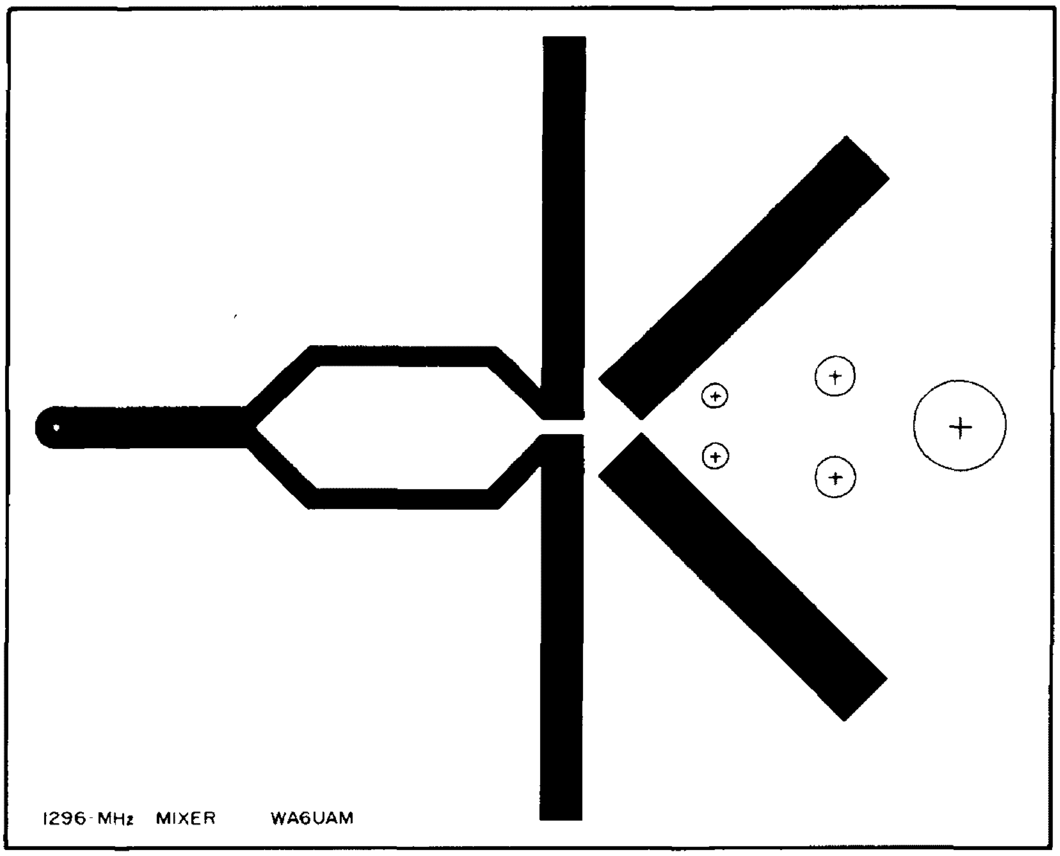


fig. 8. Full-size printed-circuit layout for the balanced mixer of fig. 5. Material is 1/16" (1.5 mm) thick, double-sided Teflon-fiberglass circuit board.

Physically, the bandpass filter can be a half- or quarter-wavelength coaxial resonator, or a trough-line cavity such as has been used in previous 1296-MHz receiving converters. Coupling in and out can be accomplished by means of links, loops, taps, capacitors or even the coaxial matching scheme used by K6UQH in his latest converters.[5] In the interest of avoiding multiple responses, a similar

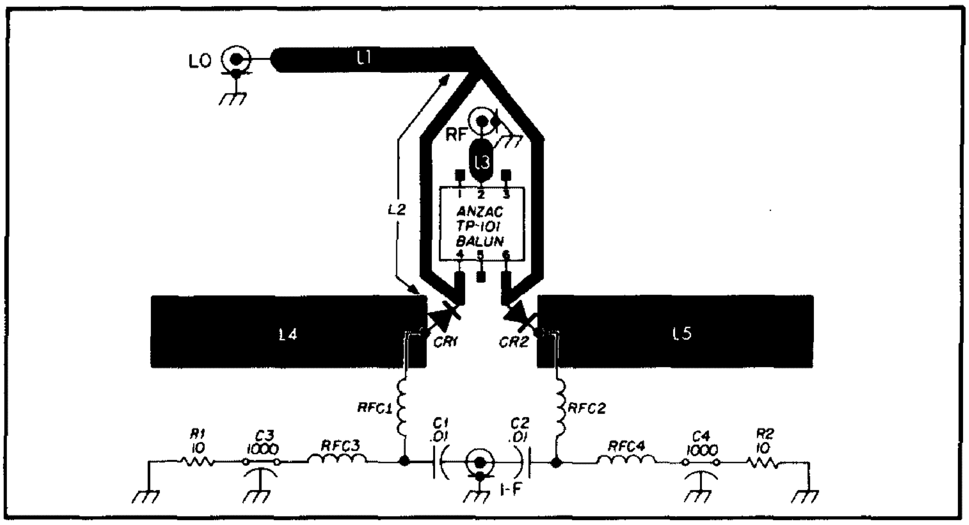
filter at the output of the LO chain may prove desirable.

local-oscillator chain

The key criteria here are stability and spectral purity. For maximum stability it is advisable to invest in the best possible low-temperature coefficient crystal you can afford.* An additional ten dollars invested in a quality crystal can do much to alleviate slight frequency drift which (when multiplied into the microwave region) can make ssb transmission and reception a running battle between your right hand and the tuning dial. The oscillator circuit should be designed to provide minimum crystal feedback consistent with ready starting, and should of course be buffered.

Spectral considerations dictate very careful selection of the multiplication scheme used to reach the desired injection frequency. High-order multiplication in a single stage is out, as the resultant harmonic comb requires extensive filtering. W6FZJ, whose success on both 432-MHz EME and 2304-MHz troposcatter speaks well for his expertise in such matters, recommends doubling in every stage of any LO multiplier. If the starting frequency (the crystal-oscillator stage) is in the 70- to 125-MHz region, any spurious responses present after repeated doubling will be sufficiently separated in frequency to be easily filtered.

The balanced mixer used at WA6UAM receives 40-mW of LO injection. Obtaining this power level at 1200 MHz is relatively simple with today's low-cost



C1,C2	0.01 μF disc ceramic	L4,L5	25-ohm micro-stripline, 0.30" (7.5 mm) wide, 1.14" (29 mm) long
C3,C4	1000-pF feedthrough	R1,R2	10-ohm, ¼-watt carbon composition resistor
CR1, CR2	hot-carrier diodes (H-P 5082-2818)	RFC1, RFC2	2" (51 mm) no. 32 wire, close wound on 0.050" (1.5 mm) form or ferrite beads on leads of C1 and C2
L1,L3	50-ohm micro-stripline, 0.10" (2.5 mm) wide, any length	RFC3, RFC4	22 μH
L2	75-ohm micro-stripline, 0.045" (1.0 mm) wide, 1.24" (31.5 mm) long (along center)		

fig. 3. Improved 1296-MHz balanced mixer uses commercial balun and double-clad epoxy circuit board. Full-size printed-circuit layout is provided in fig. 10.

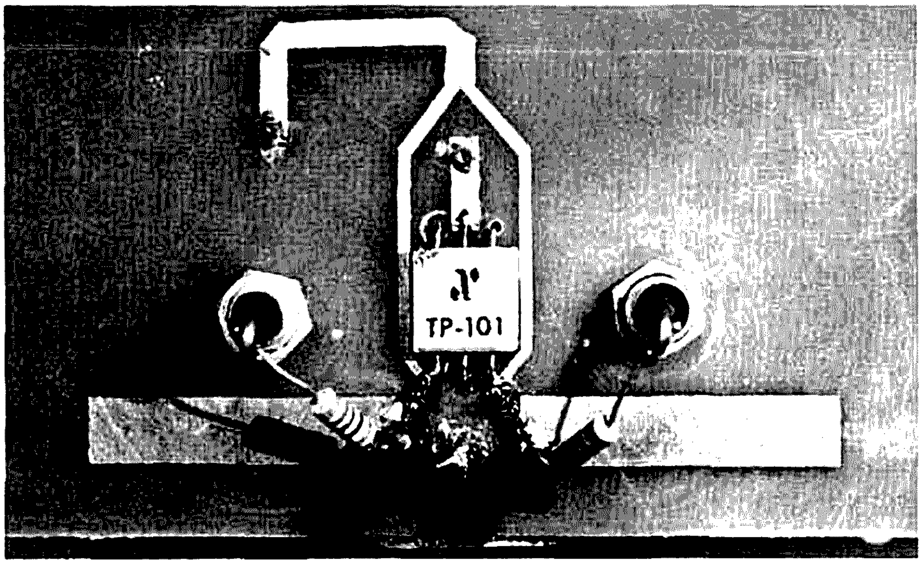
*The Croven C180DBX-00 5th-overtone crystal is highly recommended. This series-resonant, HC-18/U crystal has a calibration tolerance of ±10 ppm and temperature tolerance of ±5 ppm from 15° to 35° C. $12.00 in single quantities, or $8.00 each for two or more crystals of the same frequency. Write to Croven Ltd., 500 Beech Street, Whitby, Ontario, Canada.

uhf transistors. Stability will be enhanced by designing the second stage of the LO chain (buffer) for 100-mW output, and running succeeding doublers at or close to unity gain. Resistive 3-dB pads between stages will alleviate excessive drive and provide impedance matching.

When designing active frequency doublers, recall that second-harmonic generation is enhanced by a collector conduction angle of 180°. This condition is easily met by grounded-base, zero-bias operation. Fortunately, this is also probably the simplest frequency-multiplier circuit to drive and adjust. Incidentally, it was stated earlier that the components of the Simple Sideband System could be duplicated without the use of specialized test equipment. In the case of the LO chain, alignment can be accomplished and injection measured by merely monitoring the diode current of the balanced mixer.

local-oscillator circuits

The primary advantage of the modular system which I use is the ease with which substitutions can be made, and performance of various circuits compared. The Simple Sideband System has already worn three different LO chains; no doubt others will be attempted in the future. The block diagram of fig. 12 shows what



Construction of the doubly-balanced mixer using the Anzac TP-101 balun.

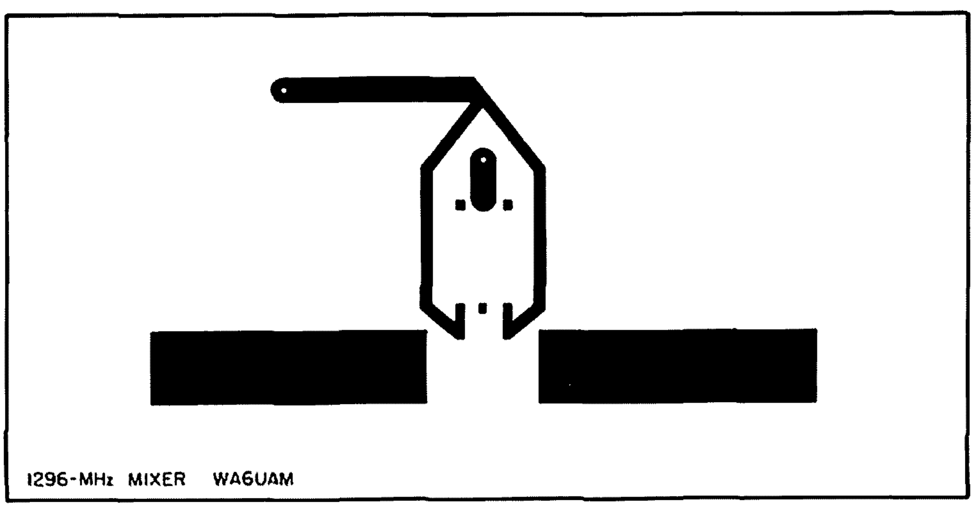


fig. 10. Full-size printed-circuit layout for the improved 1296-MHz mixer. Board is 1/16" (1.5 mm) thick, double-clad 1-ounce copper G-10 fiberglass-epoxy circuit board.

has thus far proved the most workable compromise between stability and spectral purity on the one hand, and low cost, simplicity and ease of alignment on the other.

However, this LO chain violates a number of the *ideal* design principles outlined above. For example, the purist will want the last multiplier to double rather than triple, and would probably use an active device rather than a diode. The decision to go with a diode multiplier was based primarily on the cost of 1.3 GHz transistors at the required power level. Tripling was used here because of the greater ease of generating a high level of power at 422 MHz, as compared to 630 MHz. The common-base configuration, although simple, was abandoned in the lower-level stages in deference to the greater gains available from common-emitter circuits.

The active multipliers are all operated at a low power level as a concession to stability, with the two 422-MHz power amplifiers providing plenty of drive to the tripler. Considering the low conversion efficiency of the diode tripler circuit, a high drive level is a must.

The low-level stages shown in fig. 13 were designed by W6FZJ for use in a 432-MHz receiving converter, and published in his *432 Newsletter*. Except for the crystal frequency and the number of turns on two inductors, his original circuit is unchanged when used in the 1296-MHz LO chain. Copies of the original circuit have been successfully built by a number of San Francisco area 432-MHz enthusiasts, and spectrum analyzer tests prove the W6FZJ design to be superior to any of my own attempts to date.

Since the tripler which raises the injection frequency to 1267 MHz exhibits about 10-dB conversion loss,* a half-watt of drive is necessary at 422 MHz to achieve the desired 40-mW of LO injection. This is accomplished by applying the 10-mW output of the low-level LO module to two stages of power amplification, operating at 10- and 7-dB of gain, respectively. The 2N3866s were selected because of their low cost and ready

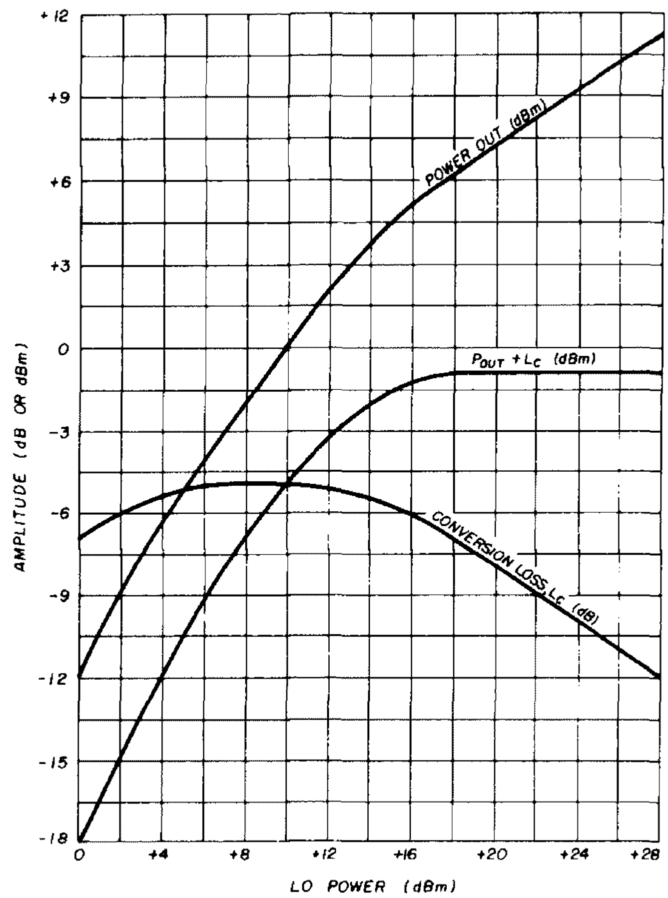


fig. 11. Balanced mixer conversion loss, output power at the 1-dB compression point and their sum (transceive figure of merit) for a typical microwave balanced mixer, all as a function of local-oscillator injection level. The transceive figure of merit represents the i-f output power available from a receive mixer driven directly by the output of an identical transmit mixer.

availability from a number of mail-order surplus component dealers.

The power amplifier circuits are shown in fig. 14. Care should be taken to closely duplicate the input and output tank circuits unless a spectrum analyzer is available, as adequate spectral purity occurs when these particular circuits are tuned for maximum indicated output.

The slab inductors in the collector

*Greater tripler efficiency is possible through the use of an idler circuit, a technique avoided by the author in the interest of simplicity.

circuits provide high output Q, and the pi-network filter (C8, L3, C9) feeding the second stage tends to suppress any harmonics generated by the first. The first 2N3866 in biased to class AB for increased gain; the second stage is run in class C for improved collector efficiency. If the stages are built separately, each can be tuned for maximum power into a 50-ohm load.

The diode tripler and filter assembly was first built in the popular trough-line configuration. Later, an interdigital filter was attempted. In both cases performance was satisfactory. However, the construction process required extensive metalworking. As many amateurs avoid projects which involve bending sheet metal or cutting brass tubing, I decided to reduce the tripler/filter to a PC board. The result is shown schematically in fig. 15.

Rf energy from the 422-MHz power amplifier, fig. 15, is applied to CR1, a GE 1N4154 high-speed switching diode through an L-network (L1, C1) similar to that used by K6UQH in his trough-line multiplier/mixers. The harmonic comb developed at the output of the diode is applied to a two-pole resonator (L2, C2, L3, C3) which blocks all but the 1267-MHz component.

The filter combines lumped and distributed constants. Design of the poles was accomplished as follows: It was desired that the filter resonate with the tuning capacitors (C2, C3) at midrange, or 1.5 pF. Assuming an additional 0.5 pF of stray and coupling capacitance, the capacitive reactance of 2 pF at 1267 MHz is 62.8 ohms. To resonate this circuit the inductors (L2, L3) must exhibit the same reactance at the frequency of interest. A shorted micro-stripline can be used as an inductor, its reactance determined both by characteristic impedance (width) and phase angle (length) as given by the relationship

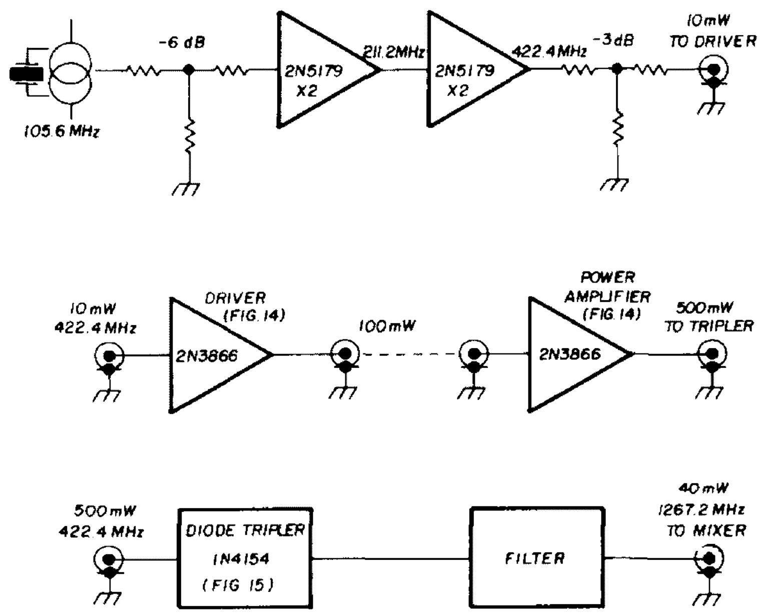


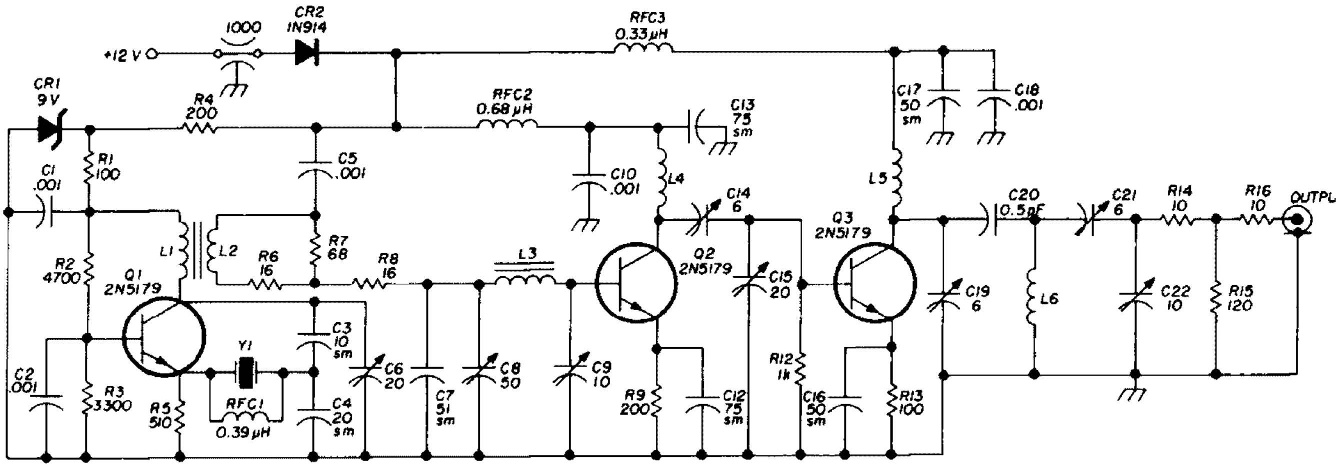
fig. 12. Block diagram of the local-oscillator chain used with the Simple Sideband System for 1296 MHz. Each of the individual sections is built into a small module; these are connected together with miniature 50-ohm coaxial cable.

$$\theta = \arctan \frac{X}{Z_o}$$

where θ represents wavelength of the stripline in *degrees*. To convert to fractions of a wavelength, divide θ by 360°. Thus, for 62.8-ohm inductive reactance with an arbitrarily selected stripline characteristic impedance of 25 ohms,

$$\theta \arctan \frac{62.8}{25} = 68.3^\circ = 0.19 \text{ wavelength}$$

A 25-ohm micro-stripline on 1/16-inch (1.5-mm) G10 PC board is 0.3-inch (7.5-mm) wide. The 0.19 wavelength at 1267 MHz (correcting both for velocity factor and width-to-height ratio) is 0.865 inch (22 mm).



C6,C15	20-pF trimmer (JFD DJV 300)
C8	50-pF trimmer (JFD DVJ 305)
C9,C22	10-pF trimmer (JFD DVJ 302)
C14,C19, C21	6-pF trimmer (JFD DVJ 301)
C20	0.5-pF ceramic (two 1-pF capacitors in series)
L1	10 turns no. 28 on Micrometals T-25-13 toroid core
L2	4 turns no. 28 on cold end of L1
L3	12 turns no. 28 on Micrometals T-25-13 toroid core
L4	6 turns no. 24, close wound on 0.1" (2.5 mm) diameter form
L5,L6	1½ turn no. 22, 1/8" (3 mm) diameter, 1/4" (6 mm) long
RFC1	0.39 µH miniature inductor (Nytronics Deci-Ductor)
Y1	5th overtone crystal, series resonant, HC-18/U holder, 105.600 MHz for 28.8-MHz i-f

fig. 13. Crystal-controlled local oscillator circuit, based on a design by Joe Reisert, W6FZJ.

Matching the resonators to the relatively low impedances of the diode and the output transmission line can be accomplished by tapping up on the micro-striplines the required distance above ground. Although formulas exist for approximating the required tap position, matching in the circuit shown was determined empirically.

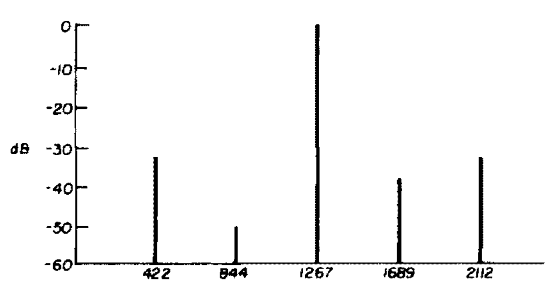
An important (and often neglected) consideration in diode multipliers is bias current. Resistor R1 in **fig. 15** enables the diode current to be varied over a wide range. Remember that diode current will affect conduction angle, which should be 120° to maximize third-harmonic generation.

transmit i-f attenuator

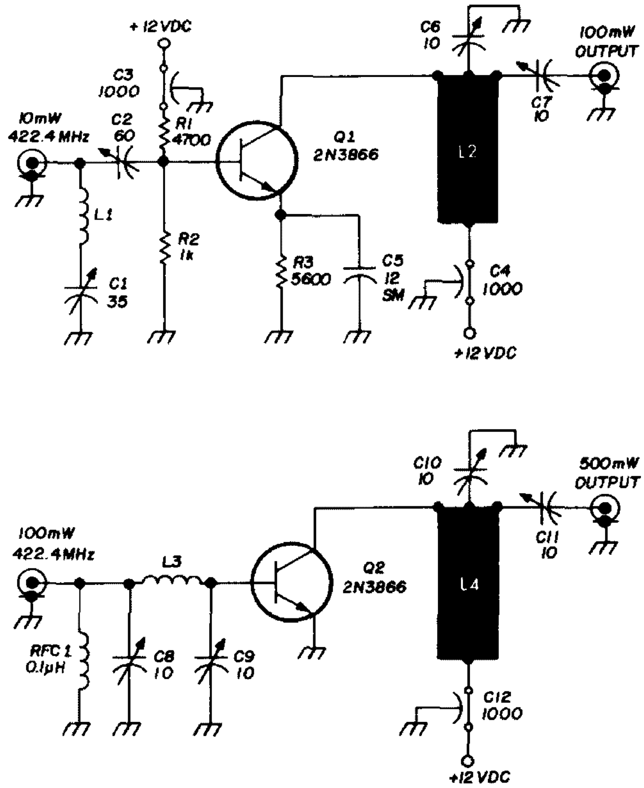
Proper operation of balanced mixers requires that each port be terminated in its characteristic impedance (usually 50 ohms). Most methods used to sample a low-level ssb signal from a high-frequency transmitter would result in a horrendous impedance mismatch at the mixer's i-f port. As 12.6-mW of sideband injection is desired, one recommended method of assuring i-f impedance matching is to run about one watt of ssb into a 20-dB resistive pi- or T-pad at the mixer's i-f port. Since the attenuation pad will provide the proper termination to the mixer, the method of coupling out of the station transmitter has no effect on the mixer's operation.

i-f amplifier

The noise figure of a receiving converter without pre-conversion gain is the



Spectrum output of the 1267.2-MHz local-oscillator chain when tuned for maximum mixer diode current. The 422- and 2112-MHz spurs are down 32 dB, the 1689-MHz spur is down 38 dB and the 844-MHz spur is down 50 dB.



C1	3-35 pF trimmer
C2	8-60 pF trimmer
C6-C11	10-pF concentric piston trimmers
L1	2 turns no. 18, wound on 1/4" (6 mm) mandrel, 1/8" (3 mm) long
L2,L4	brass strip, 0.5" (12.5 mm) wide, 1.5" (38 mm) long, mounted 1/8" (3 mm) above ground plane
L3	2 turns 1/8" (3 mm) wide brass strip, 0.1" (2.5 mm) diameter, 0.5" (12.5 mm) long

fig. 14. The two 422-MHz power amplifiers provide one-half watt output. The first stage, Q1, provides 10-dB gain, while the second stage, Q2, provides 7-dB gain.

sum of the feedline and TR relay losses (if any), input filter insertion loss, mixer conversion loss and i-f amplifier noise figure. For the system presented here the noise contribution of these stages prior to the i-f amplifier is less than 6.5 dB. Feeding the output of the mixer into a high-frequency ssb receiver would, however, result in a system noise figure approaching 20 dB. Many amateurs are surprised to learn that the noise figure of even a high quality commercial communications receiver is seldom below 10 or 15 dB. In the high-frequency spectrum the level of man-made and atmospheric noise exceeds that of receiver noise by several orders of magnitude, so noise figure, per se, is not usually a significant consideration in high-frequency receiver design. Of course, such is not the case in the microwave region. To achieve a reasonable noise figure in any uhf converter, a low-noise i-f amplifier must be used to mask the following receiver noise.

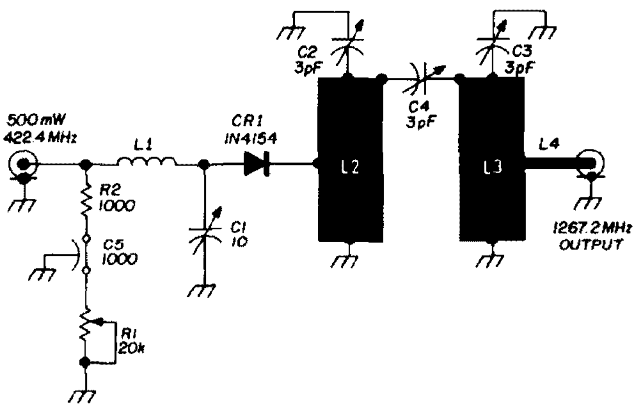
Numerous circuits have been described in the past which will yield reasonable gain at 28 MHz with a noise figure under 1 dB. WB6NMT, known for his pioneering efforts in 220-MHz EME, uses neutralized fets, while W6FZJ favors a dual-gate mosfet following the mixer. An appealing circuit by W9PRZ uses a dual jfet in a cascode configuration and is described in a Siliconix applications note.[6]

system performance calculations

It remains to be shown that the simple microwave ssb system described here lends itself to satisfactory communications over reasonable distances (my experience shows the following calculations to be somewhat on the conservative side). The overall noise figure of the receive system is the sum of filter loss (0.5 dB), mixer conversion loss (6 dB) and i-f noise figure (1 dB). Thus, a receive noise figure of 7.5 dB is assumed. With the 2.1-kHz i-f bandpass of a good sideband receiver, the graph of fig. 16 indicates the receive sensitivity to be about −133 dBm.

Good intelligibility on single sideband requires a 6-dB signal-to-noise ratio. Allowing a 10-dB signal-to-noise ratio for good measure, the receiver requires a −123 dB signal. At +4.3 dBm of transmitter power, given 10-dB of antenna gain at each end, and allowing 5-dB of additional loss for antenna aiming errors, the permissible path attenuation between transmit and receive stations is 142.3 dB. The free-space loss formula is[7]

$$L_{FS} = 36.6 + 20 \log_{10} f_{MHz} + 20 \log_{10} d_{miles}$$



C1	1-10 pF concentric piston trimmer
C2,C3 C4	0.3-3 pF concentric piston trimmer
CR1	1N4154 high-speed switching diode
L1	2 turns no. 20, 0.1" (2.5 mm) diameter, 0.25" (6 mm) long
L2	micro-stripline, 0.3" (7.5 mm) wide, 0.865" (22 mm) long, grounded at bottom, tapped 0.20" (5 mm) from ground end
L3	Same as L2 but tapped 0.25" (6 mm) from ground end
L4	50-ohm micro-stripline, 0.1" (2.5 mm) wide, any length
R1	20k, 10-turn trimpot

fig. 15. The 422- to 1267-MHz diode tripler is built on 1/16" (1.5 mm) double-clad, glass epoxy circuit board.

which can be rewritten to solve for distance at 1296 MHz

$$\log_{10} d_{miles} = (L_{FS} - 98.85)/20$$

Given 142.3 dB of permissible free-space loss, an HP-35 calculator yields

$$d(\text{maximum}) = 161 \text{ miles } (259 \text{ km})$$

Amateurs who have operated simple equipment in the 23-cm band find it difficult to believe that a 161-mile path is possible with only 3-mW of output power. After all, they reason, that's greater than the APX-6's range, and those have 3-watts output. It should be remembered, however, that directly-modulated oscillators tend to be extremely unstable, and that reception of their emissions requires a wideband receiver. The APX-6 i-f strip is about 5-MHz wide and bandwidth has a serious impact on receiver sensitivity. **Fig. 16** indicates a deterioration in receiver sensitivity of 10 dBm for each tenfold increase in i-f bandwidth. This more than offsets any path increase afforded by the greater power of the APX-6 or similar equipment.

direct conversion

If simplicity is the goal, there is practically no limit to the evolutionary process. This last step occurred almost by accident, while developing test equipment for 1296 MHz. A weak-signal source was built for 1296.00 MHz to permit calibration and receiver testing. Since a few milliwatts of stable power was desired, I used the same techniques as I used for building the converter local oscillator. Later a key was added, and the signal source used for on-the-air CW contacts over a limited range.

Eventually, a means for monitoring transmit quality from the ssb converter was desired. An antenna and tuned cavity feeding a diode detector and audio amplifier produced the characteristic ssb "Donald Duck" squawk in a speaker. However, for high-quality signal monitoring a ssb converter was required. By using the 1296.0-MHz signal source developed earlier, driving a balanced mixer with its i-f port feeding an audio amplifier, a direct-conversion receiver was built. If the

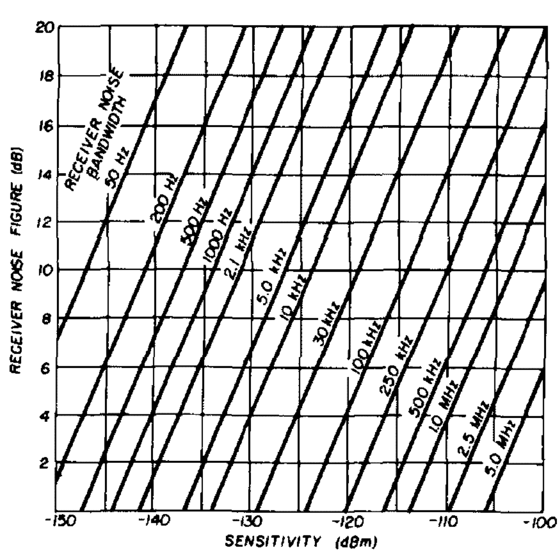


fig. 16. Receiver noise figure vs sensitivity for various receiver bandwidths.

transmitter (suppressed) carrier frequency and the receiver LO frequency are the same, the difference frequency at the i-f port of the mixer will be audio sidebands. No input filter is required because the image is merely the other sideband – and at 1296 MHz QRM is hardly a problem yet.

If a direct-conversion double-sideband receiver will work, there's no reason to expect otherwise for direct double-sideband generation. Indeed, when audio was applied at the i-f of the balanced mixer, sidebands around 1296.0 MHz appeared at the rf port. Though extremely low power, the signals could be copied in a sensitive receiver at a range of about a mile.

The resulting direct-conversion double-sideband transceiver is shown in **fig. 17**. Though primarily a lab accessory and demonstration rig, it provides reliable communications over moderate distances. It is simple in the extreme, and suggests the possibility of microwave double-sideband walkie-talkies. The concept may even work at greater power levels than the few microwatts attempted to date. Certainly there's no more basic a way to produce a microwave sideband station.

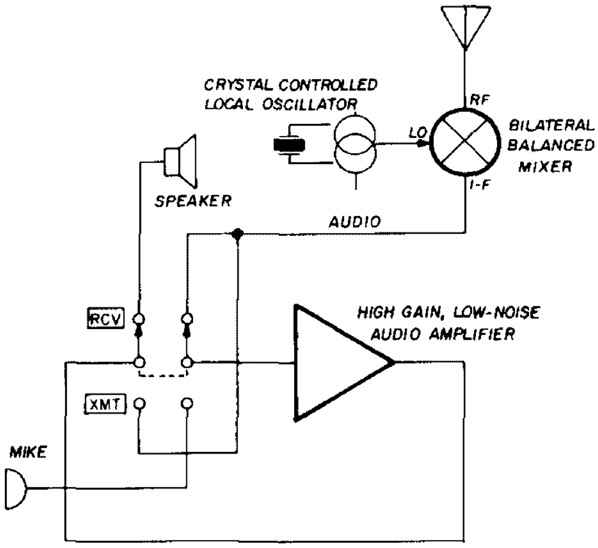


fig. 17. Simple direct-conversion double-sideband transceiver for 1296 MHz.

conclusion

The primary advantage of ssb in the microwave spectrum, as anywhere else, is in the significant increase in receiver sensitivity resulting from the narrow bandwidth which is required. The only limiting factor is stability, a criterion which can be readily achieved by judicious application of good engineering practice to the design and construction of local-oscillator chains. The design tradeoffs presented here make microwave ssb feasible over considerable distances, while requiring a minimum of specialized skills, equipment or technique.

acknowledgements

I wish to express my deep appreciation to Don Farwell, WA6GYD, for giving me my first contact on 1296 MHz and kindling the fire, to Frank Pacier, W6VMY, for the countless hours we spent together working on APX-6s in the early days, to Bill Troetschel, K6UQH, who shamed me into giving up modulated oscillators by his snide remarks at uhf conferences, to Bob Ney, WB6LLD, who gave me my first microwave transistors, and to Joe Reisert, W6FZJ, who taught me how to use them. And most important, a word of thanks to my wife, Suk, who for the past three years has shown more understanding of the hours spent on the 1296 project than I had any right to expect.

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ham radio

miniaturized communications receiver

Full construction details for the Minicom — a complete communications receiver that can be used on the 40- and 80-meter bands

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

This article has to do with the art of homebrewing your own communications gear — the specific application is a miniature communications receiver. It was inspired by the huge packet of mail I received following publication of my article extolling the virtues of the LM373 IC manufactured by National Semiconductor.[1] Most of the letters contained requests for additional construction details, parts information and printed-circuit artwork. Unfortunately, the original receiver contained some odd-ball components, and the board layout went the way of all trash during the year that passed between submission and publication, so I had to play the heavy and turn down most requests. As time went on and similar letters arrived, even a year later, I decided to do it all over again as a real construction article.

the homebrew art

If you are planning to skip this part, do so at your own risk because I may throw in a few clinkers during the construction phase that will cause untold agony for those who skip ahead.

Obtaining needed parts is probably the most annoying problem plaguing the do-it-yourself fan. Having an up-to-date library of standard and surplus catalogs does not even guarantee success. Even a calm, logical, well-ordered person can be reduced to a shattered, babbling hulk as a result of unsuccessful forays into the electronics marketplace. That's probably why kits have become so popular. For the intrepid craftsman, however, this may not be a practical answer since the manufacturers don't always kit what you want to build. To overcome these roadblocks, ingenuity and improvisation become the key factors in maintaining mental health

and ensuring completion of a cherished project.

Aside from improvising in the parts department, miniaturization is generally best implemented by using printed-circuit construction. It is in this area that thoughtful layout is all important. Without carrying the shrinking process too far, a nicely balanced design has been achieved in this receiver without overcrowding or requiring fussy assembly techniques. The principal factor that limits size in this project is panel space. There is no sense in shrinking the receiver to postage stamp size if you need ten times that much space for the essential front-panel controls.

Integrated circuits are now a way of life and often provide the only means by which ten pounds of circuitry can be squeezed into a five-pound bag. This advantage can be lost, however, if an IC chosen for a particular function requires more external components than the same circuit would use in discreet form.

On top of this, you become a prime candidate for the funny farm by virtue of spending three days trying to lay out the artwork for this nightmare. By using a little discretion, ICs can be selected which will provide all the performance you expect while requiring a minimum amount of valuable printed-circuit real estate.

printed circuits

The above comments are of little use if your etchings are a lost cause. Of course, not all do-it-yourselfers use PC assemblies, but these techniques do make a project more professional in appearance and simplify duplication. Though most of the methods suitable for home fabrication of PC boards have been thoroughly covered in previous articles, I'd like to throw in a few comments on my method since I find it simpler, faster and superior to any of the others.

First, the circuit is laid out in pencil to exact scale on one-tenth (25 mm) grid paper. You'll have to develop your own skill in this phase of the operation but be prepared for a few frustrations until you get the hang of it. I try to avoid the need for jumpers but this is not always possible.

The next step is to lay out the positive of the printed circuit with sticky-back circuit trace tape and pads commonly used for this purpose. If there are no copper areas other than the lines and circles, a piece of clear film may be placed over the drawing and the tape and donut-shaped pads applied to the transparent film. If copper areas, such as ground planes, are present in the design, as they are in this receiver, a piece of Rubylith† is more appropriate. This material consists of a clear polyester base on one surface of which is a thin red film. Sections of this film may be stripped away after first scoring the perimeter with a sharp knife. The remaining circuitry is applied with tape and pads as before.

When the positive artwork has been completed, a negative is prepared by contact-printing the artwork on a piece of 3M reversing film. A 30-second exposure in bright sunlight followed by a few seconds for developing result in a negative replica of the original artwork. This negative is then used to contact-print the circuit on a piece of sensitized copper-clad board. Here again a 30-second shot of sunlight will do it. The circuit board is then developed and etched in the normal manner.

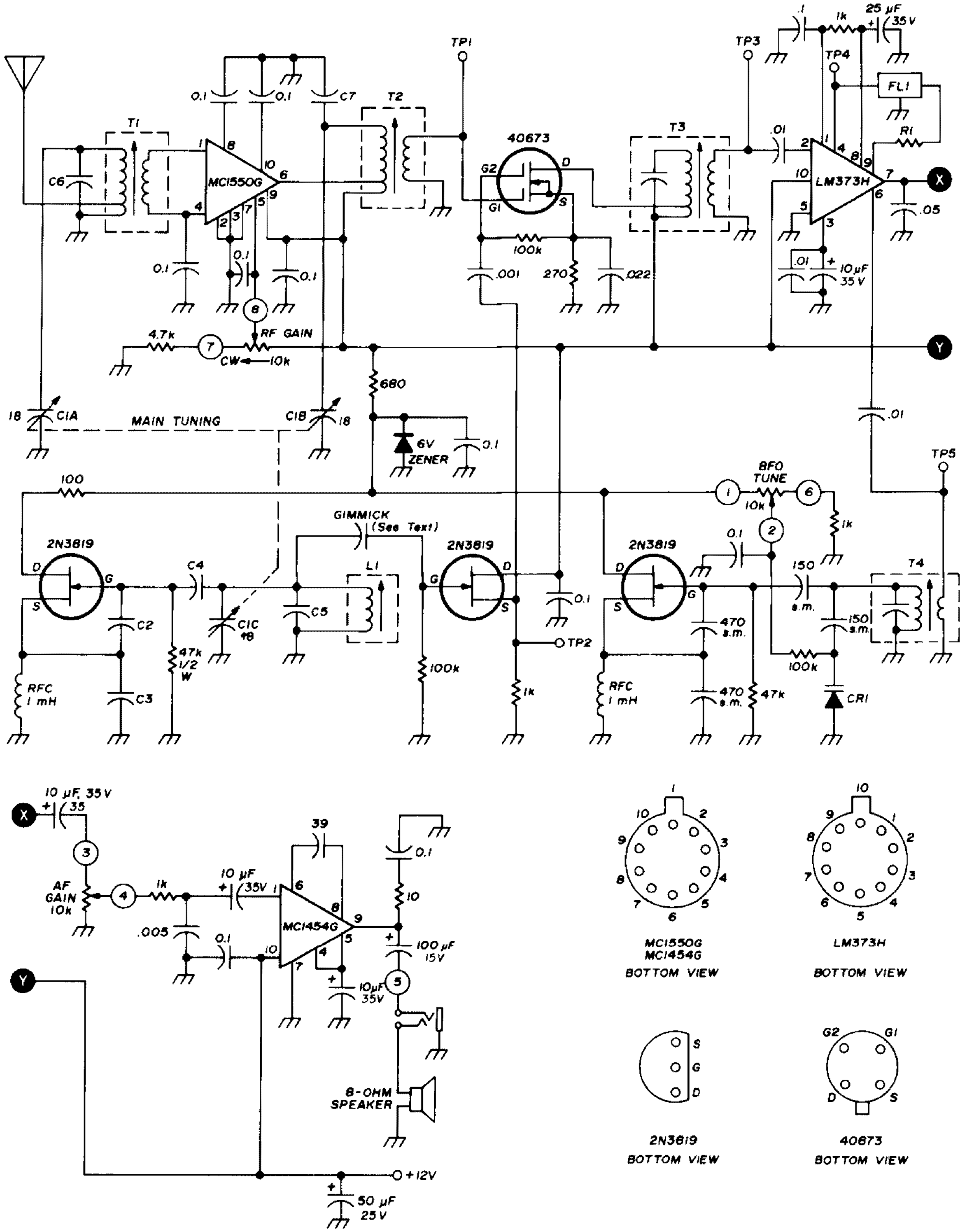
The 3M reversing film is a member of the Scotchcal* family which also includes sensitized aluminum sheet stock suitable for making panels, name plates, dials, etc.

The time required for preparing the original artwork may vary from a matter of minutes to several days, but once complete, this method will produce a finished circuit board in a matter of minutes, even if you include the time spent in coating the board with photoresist. No camera work or darkroom is required and the results are as good as the original layout.

Incidentally, a simple method for

†Rubylith is a registered trademark of Ulano.

*Registered trademark of the 3M Company.



C1	3-section variable, 18-pF per section (J.W. Miller 1460 or similar)	FL1	Ceramic filter (Murata SFD455D or CFS455J, see text) (a 56-pF coupling capacitor is used with the SFD455D and is not shown in the schematic)
C2,C3	220-pF for 80, 470 pF for 40 (silver mica)	L1	see fig. 2
C4	39 pF for 80, 100 pF for 40 (silver mica)	R1	3000 ohms for SFD455D filter, 2000 ohms for CFS455J filter
C5	15 pF for 80, 70 pF for 40 (silver mica)	T1,T2,T4	see fig. 2
C6,C7	20 pF for 80, 130 pF for 40 (silver mica)	T3	miniature 455-kHz if i-f transformer

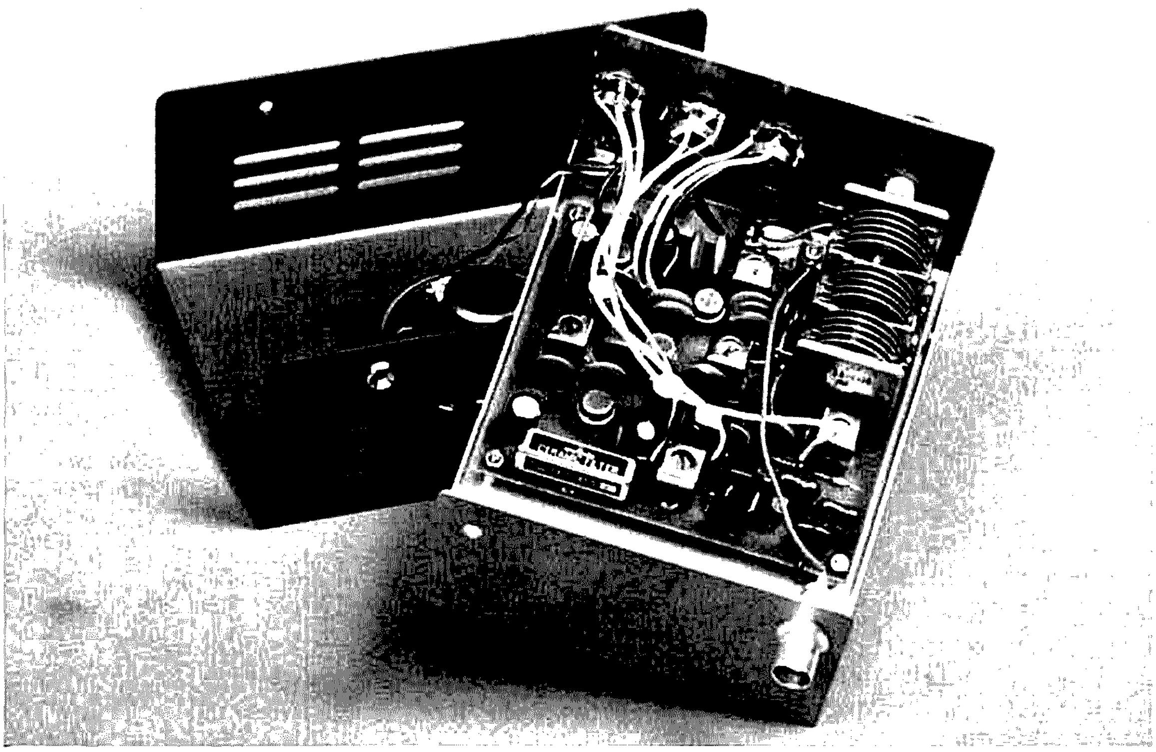
fig. 1. Schematic for the complete minicom receiver. Circled numbers correspond to the numbers on the printed-circuit layout (fig. 3).

achieving a rapid etch in ferric chloride is to brush the copper surface constantly with a soft paint brush while the board is immersed in the solution. One-ounce copper in fresh solution will etch in four or five minutes using this method while simple agitation may take four or five times as long.

druggist. For those of you who want to buy everything in small quantities from one place, Kepro packages the above materials and sells through several of the large mail-order houses.

minicom receiver

The miniature communications re-



Inside the cabinet of an earlier version of the receiver which used an IC mixer. This copy also sports the more selective ladder filter seen along the left rear edge of the board.

Another point worth mentioning is that when coating your copper stock with photoresist, a thin layer is always good insurance against problems during the developing process. This is especially true if trichloroethylene is used as the developer. I paint the resist on with a cotton swab and avoid spray cans like the plague.

I buy my Rubylith, reversing film and photoresist from a photo supply house. Trichloroethylene can sometimes be found in drug or hardware stores but I find it less expensive to buy a gallon (3.8 liters) from a local chemical supply agency. Developer for the reversing film is that ever-popular number, isopropyl alchohol, which is available from your local

ceiver described here is similar to the one I mentioned earlier. The only significant change is that an IC is used in place of an fet in the rf amplifier (see fig. 1). The original rf stage had a tendency to oscillate and the MC1550G IC completely eliminates this problem. An rf gain control was also added since this is easily accomplished with the IC.

In some versions of the receiver I experimented with, I used an IC balanced mixer. The MC1496 was considered but was abandoned because of the relatively large number of external components it required. Instead, an SG3402T by Silicon General was used since it only requires a couple of bypass capacitors as external components. This design worked fine and

one of the receivers pictured here has the IC mixer.

The other receiver has a dual-gate mosfet mixer — this is the version which is described in detail. The decision to go with the mosfet mixer was purely philanthropic. Searching for an SG3402T IC might result in considerable mental anguish which can be easily avoided by using the mosfet.

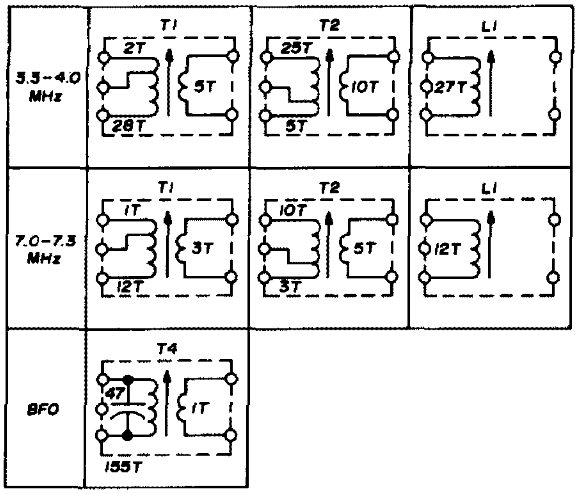


fig. 2. Construction of transformers and inductors used in the minicom receiver (see text).

The mixer output feeds the LM373H IC which performs the functions of i-f amplifier, agc system and product detector. The main selectivity-determining device is part of the LM373H circuitry and consists of a 455-kHz ceramic filter. Actually, you have a choice between a simple 2-section filter (Murata SFD455D) which results in a 3-dB bandwidth of around 4.5 kHz or a more costly ladder filter (Murata CFS455J) with a 3-kHz bandwidth at 6 dB. The circuit board is designed to accommodate either filter and the choice depends primarily on the size of your wallet. Since a receiver of this type would normally not be used in contest work or for any serious fixed-station operation, I feel the simple 2-section filter is quite adequate. Also, it costs about one tenth the price of the ladder filter.

I decided from the beginning to forget about a-m reception and to make this receiver a strictly ssb/CW affair. The receiver described in the previous article used a multigang switch to select either a-m or ssb/CW; it may be referred to if you are interested in making the required modifications. You may also wish to read that article for more details regarding the LM373H IC.

The final stage in the minicom receiver uses the MC1454G audio power amplifier IC. Here again I prefer this particular IC since it does a good job with very few external components and comes in a metal package. It will adequately drive a small speaker such as the built-in unit shown in the photograph. A larger speaker can be plugged into the phone jack.

inductors and transformers

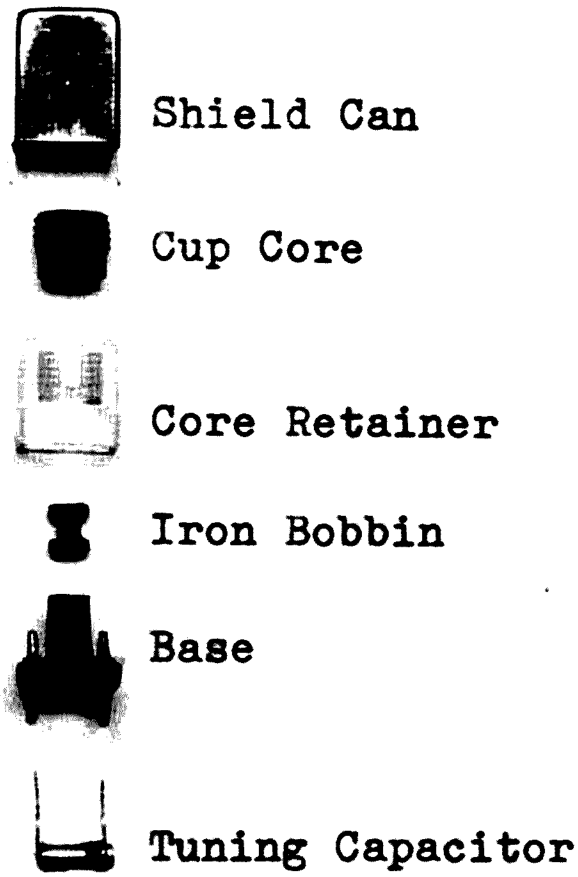
All the coils and transformers used in the construction of the receiver are fabricated from miniature 3/8-inch (95 mm) square by 1/2-inch (127 mm) high transistor i-f transformers. With a little patience and a small soldering iron, they can be stripped and rewound for so many applications that it pays to keep a drawerfull around the shack. If you've reached the bifocal stage like me, a magnifying glass is helpful.

Both 455-kHz and 10.7-MHz transformers are used for this improvisation, so stock up on both kinds. The 455-kHz transformers have a winding of about 150 turns which is resonated with a 180-pF capacitor. The 10.7-MHz fm units contain a 47-pF capacitor across about a dozen turns. The entire assembly consists of a molded plastic base, an iron core, a threaded cup core, a plastic retainer for the cup core and the metal shield can.

Most types have two small dimples alongside either solder tab on the shield can which lock everything together. Forcing a thin knife blade down the inside of the shield can and prying the dimples slightly outward will improve your chances of separating everything without damage. Gently pull on the base pins until the base assembly pulls out of the shield can. Normally the cup core and its retainer will remain inside the can and may be left there. Occasionally I've run into transformers in which the cup-core

retainer is part of the base assembly and remains attached when the base is removed. In this case, just remove the cup core and set it aside until you are ready to reassembly everything.

First unsolder all the leads running to the five pins. Also unsolder the leads from the tuning capacitor. The only transformers not having tuning capacitors are the oscillator transformers with red cores. Save these capacitors for possible use in special transformers since they are the only ones that will fit into the space provided.



Exploded view of a typical i-f transformer of the type used in the minicom receiver.

When the wires have been unsoldered the iron coil bobbin may be pulled out of its socket. This piece is generally held in place by contact cement that is easily removed. The wire used in these coils has solderable insulation and is very easy to use, so salvage as much as you can. Clamp a miniature alligator clip onto one end of the iron coil bobbin so you can hold it while the new winding is applied. If the winding has a tap, just bring out a loop at the proper turn.

If you require some center-tapped transformers, bifilar wind these using two separate wires. Put the large winding on first and squeeze some coil wax on the winding to hold it in place. Remount the bobbin and solder the leads to their proper pins. I generally don't use any cement on the bobbin, but you can if you want. The small secondary winding can now be added by soldering one end of the lead to its proper pin and feeding the wire around the bobbin for the required number of turns before soldering the remaining end.

A few well-placed chunks of coil wax are melted with a soldering iron until the wax flows in and around the coil to hold everything firmly in place when it hardens. Make sure all leads are dressed down the side of the coil and against the plastic base so there will be no interference with vertical travel of the cup core. Push the base assembly back inside the shield can and the transformer is ready to install.

Both 80- and 40-meter versions of this receiver were built around the same circuit board. The coil table in fig. 2 gives winding data and pin connections (bottom view) for all the coils. The 80-meter receiver uses 455-kHz stock while the 40-meter coils were wound on 10.7-MHz forms. A 47-pF capacitor from one of the latter is used to tune the bfo transformer. Don't forget to install this item when fabricating the bfo coil. Salvaged wire may be used for all coils except the bfo. Since you won't have a single piece of salvaged wire long enough for a 155-turn winding, use some fresh number-40 magnet wire.

construction

The PC board is 3.7-inches (9.4 cm) square and mounts all parts except the front panel controls (fig. 3). To prevent interference with panel-mounted controls, enough blank space has been left on

the leading edge of the board to allow a section 2.2- by 0.4-inches (5.6x1.0 cm) to be removed. By doing this, the board may be mounted close enough for the tuning capacitor shaft to protrude through the panel and still allow miniature pots to be mounted at board level.

for the mounting screws to allow alignment of the capacitor if necessary.

The only way you'll be able to mount the coils and transformers is to cut slits in the board for each of the mounting tabs on the shield cans. This is best done with a small pointed knife blade forced into

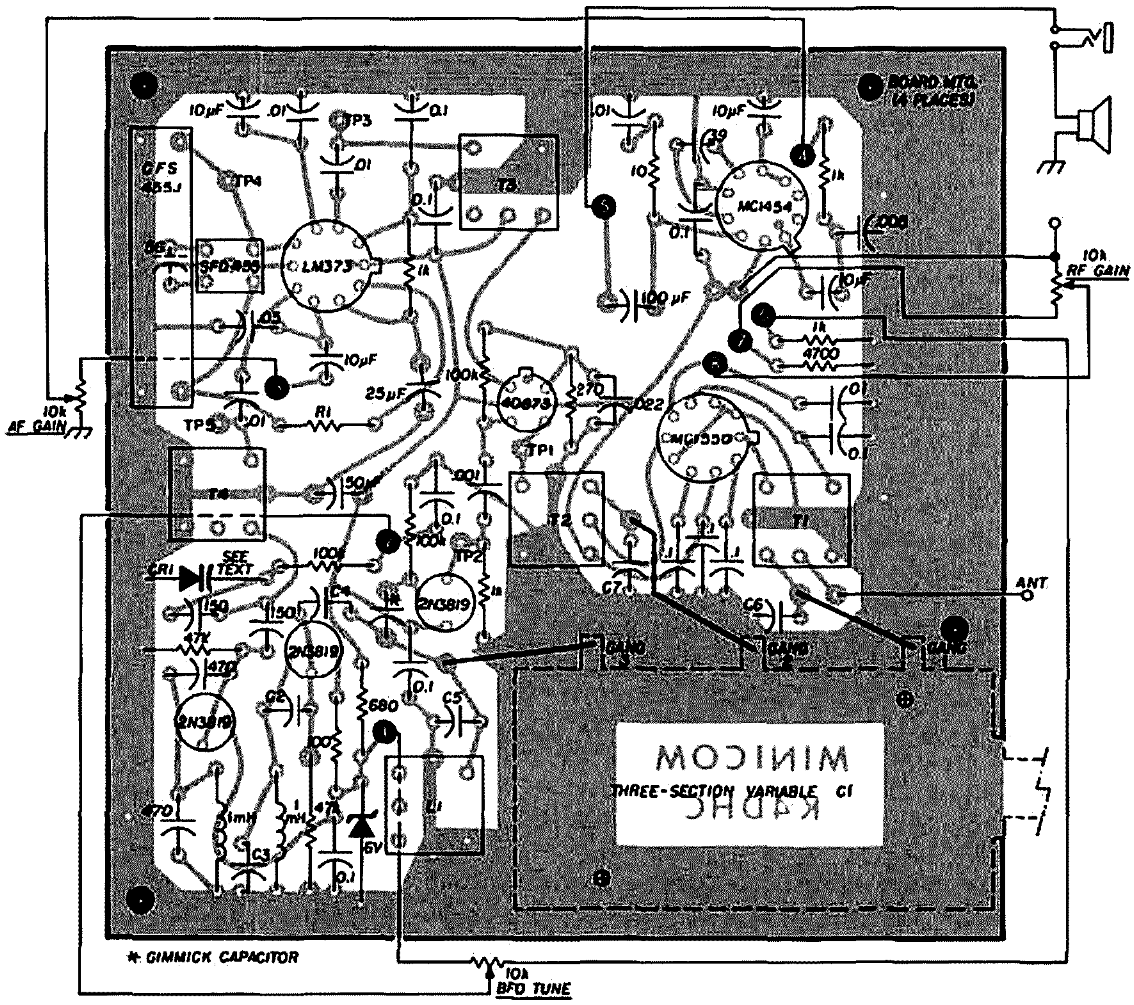


fig. 3. Full-size component layout for the minicom receiver.

When all the coils and transformers have been completed, the circuit board may be prepared and the holes drilled for components. I used a number-65 drill for the 1/4-watt resistors, chokes and all capacitors except the larger electrolytics. A number-70 drill was used for the transistors, ICs and diodes. All remaining holes were drilled with a number-60 drill. The four mounting holes may later be opened up to pass a number-2 or -4 screw. A number-30 drill was used to open up the two holes for the tuning capacitor. This provided enough clearance

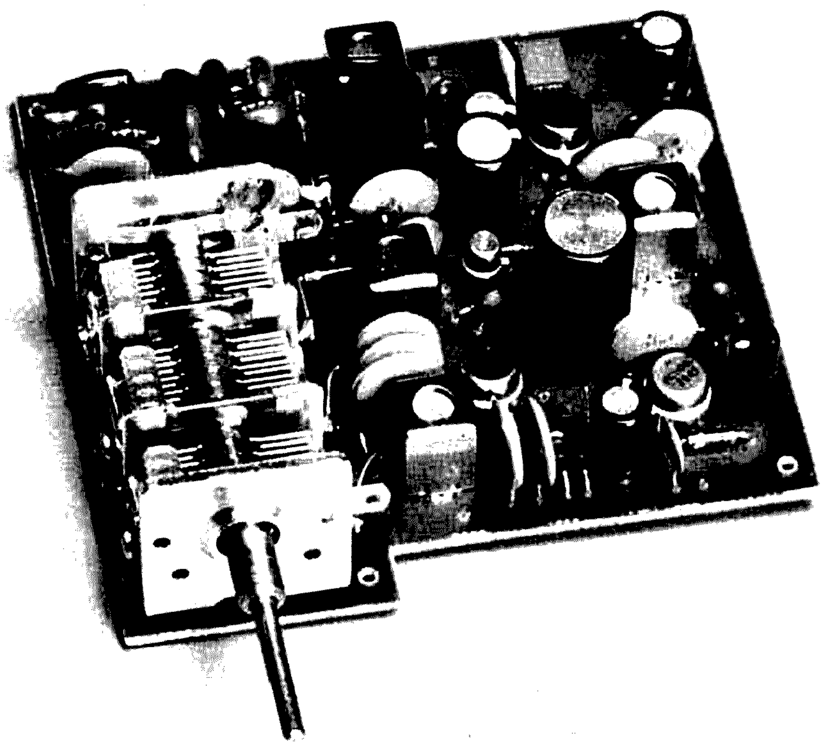
the drilled hole at the proper angle from both sides. Don't be concerned if the copper pad is cut because solder will be flowed all around the tab to hide your mechanical ineptitude. If you plan to notch the front edge of the board, do it now.

When you are ready to start mounting components, it is suggested that you use the following order which should make the process a little easier and faster. All the 1/4-watt resistors are mounted first, followed by the diodes, rf chokes and the 1/2-watt resistor. The transistors, ICs and

the filter are next, then the ceramic, mica and small electrolytic capacitors. The coils, transformers and large electrolytics are next while the tuning capacitor is saved for last.

Before mounting this last item, cut the three solder lugs down to a length of amount of coupling should be adequate but may be adjusted later if required.

Leads of regular insulated hook-up wire may be installed next for the points going to front-panel controls, antenna, the speaker or closed-circuit jack and the power supply.



The printed-circuit assembly of the minicom receiver. The mosfet mixer is almost in the center of the board. The small ceramic filter may be seen at the right rear edge of the board.

1/8-inch (32-mm). Use a flat washer between the board and capacitor frame at each mounting screw to avoid interference with the trimmer assemblies on the two forward gangs. Toothed washers under each screw head should be used to ensure good contact with the copper and prevent loosening. Connect each gang of the tuning capacitor to its appropriate point on the board with solid bus wire.

Coupling between the hfo tank and the source-follower is accomplished with a gimmick capacitor. Solder a 1/2-inch (127-mm) length of bus wire and a 3/4-inch (190-mm) long piece of insulated hook-up wire in the holes provided for the gimmick. Wrap the insulated lead loosely once around the bus wire. This

If you wish to test the receiver before installing it in a cabinet, go ahead and wire up the controls. There are two holes along the front and one along the rear edge for grounds. These are for convenience and may be used as you see fit. The use of metal spacers for mounting the board will ground the common on the board to the metal cabinet.

test and alignment

There are five test points provided on the circuit board. Each is noted in fig. 1. The padding capacitor values called out are those used in the prototype receivers, but some alteration may be necessary if tracking is poor or the frequency range is greater or less than nominal. The 80-

meter version should tune from 3.5 to 4.0 MHz and the 40-meter receiver from 7.0 to 7.3 MHz.

Power up the receiver with a 12-volt supply (fig. 4) and observe that the current with no signal is around 35 mA, indicating that you managed to avoid any

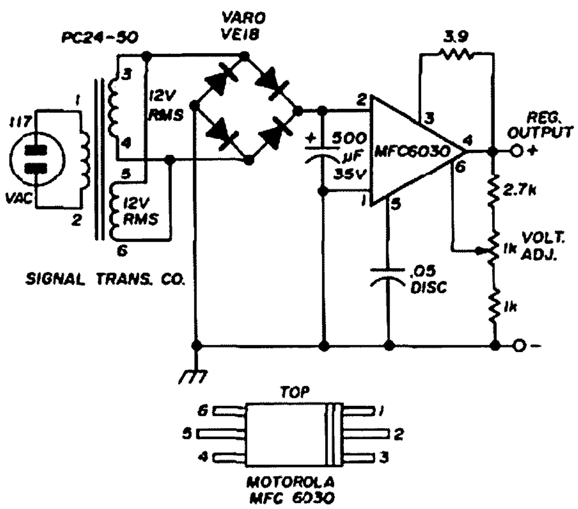


fig. 4. A 12-volt power supply suitable for use with the receiver. A printed-circuit layout for this ac power supply is shown in fig. 5.

short circuits. If smoke is absent, warm up your signal generator and start aligning. The following procedure is abbreviated, but should start you off in the right direction.

The two trimmers should be run up snug but not tight, the two gain controls turned up full and the bfo set to midrange. Adjust the core in the bfo transformer until you hear some noise, indicating the frequency is close to 455 kHz. Connect the generator to the antenna terminal and run the output up and down until you pick up a signal and know the receiver is working. Set the tuning capacitor to full mesh and the generator to the low end of the band. Adjust the core in the hfo coil until the signal is picked up.

Run the tuning capacitor all the way open and check for a signal at the high end of the band. If all is well, center up the tuning range by means of the core in the oscillator coil. Adjust the cores in transformers T1 and T2 for maximum gain at the low end of the band and the capacitive trimmers for maximum gain at the high end. Peak the i-f transformer for greatest noise. The bfo can then be adjusted so zero beat occurs about mid-position of the tuning control.

closing comments

If you don't have any capacitor diodes suitable for use in the bfo tuning circuit, glass silicon diode rectifiers generally work. That is what I used in the final version. The dipped silver-mica capacitors used in the tuned circuits should be the smallest sizes available or you may run into mounting problems. The lead spacing and outside diameter of the electrolytics should be taken into account when buying these items. Most of the ones I used were obtained from Radio Shack. Even then, it was found that capacitors marked with the same values and voltages often varied in case size. These conspiracies are all around, so be wary.

The cabinet shown in the photograph is a Radio Shack 270-252 and costs \$2.59. Dimensions are 4 x 2-3/8 x 6 inches (4.8 x 6.0 x 15.2 cm) which makes a fairly compact package. Since panel space is at a premium, the speaker and its perforated metal grill were epoxied to the top of the cabinet. There is plenty of

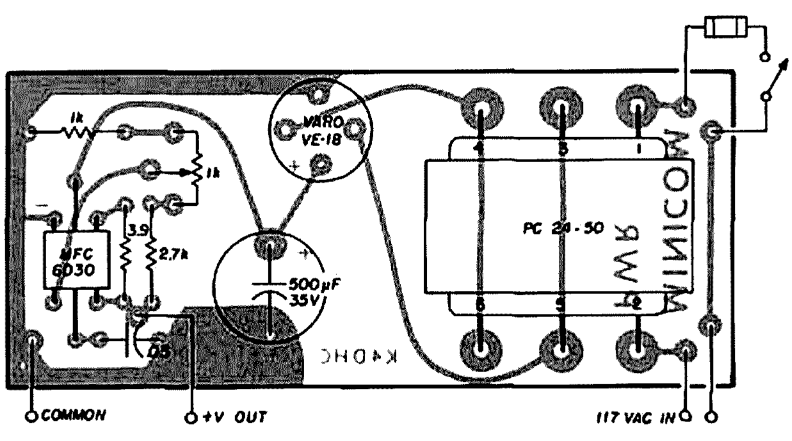


fig. 5. Full-size component layout for the ac power supply.

room left inside for a 12-volt pack of penlight cells or for a small power supply. The circuit for a power supply I built is shown in fig. 4. It will deliver about 11 volts of regulated power. The receiver will work on any voltage from about 9 to 12 volts with very little difference in performance.

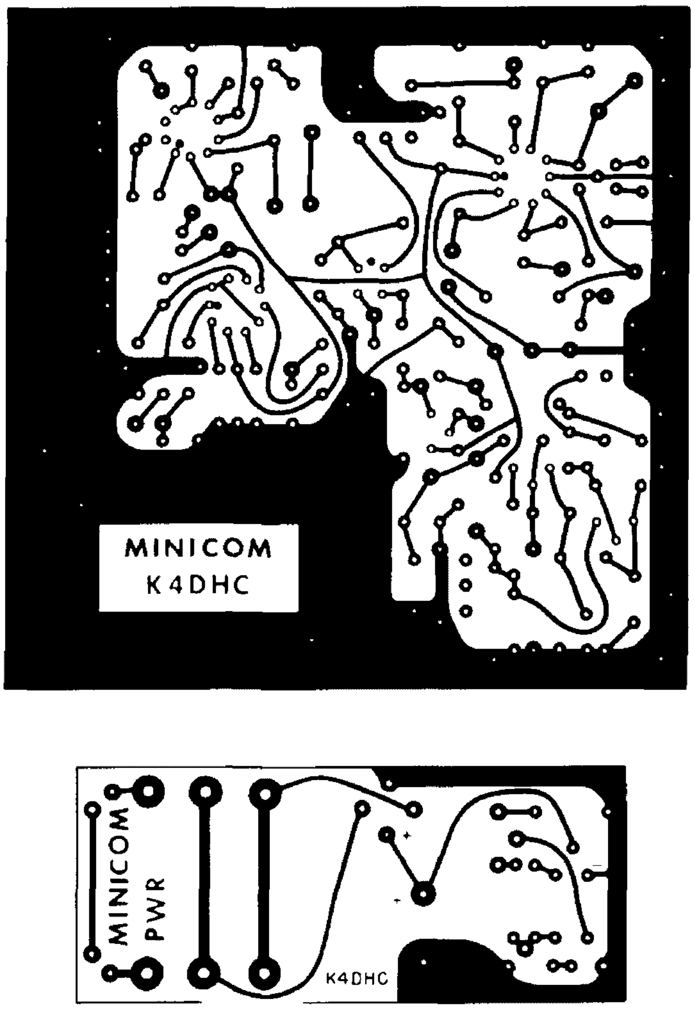


fig. 6. Printed-circuit layout for the minicom receiver and power supply (not full size).

Naturally, other variations of this receiver are possible. A two-band version which covered both 80 and 40 meters without bandswitching was breadboarded. The oscillator covered the range from 5.25 to 5.75 MHz, resulting in an i-f of 1750 kHz for both bands. The output from a crystal oscillator operating at 2.205 MHz was then mixed with the 1750 kHz signal to produce a second i-f at 455 kHz. A separate, high-capacitance two-gang tuning capacitor was used to tune the rf stage. This arrangement hit 80 meters at one end and 40 meters at the other end of the tuning range, eliminating any need for bandswitching.

Experiments with these techniques at 20 meters left a lot to be desired, however, and I found that a converter worked best for multiband operation on the higher frequencies. Odd-ball i-f transformers are a cinch, though, and many specials have been fabricated for successful use in all kinds of circuitry. The 1750-kHz i-f transformer consisted of 36 turns tuned with the original 180-pF tuning capacitor. A tap at 2 turns for the MC1550G and a secondary of 10 turns to feed the mosfet mixer completed the unit.

The stock 455-kHz oscillator transformers (red core) have no built-in tuning capacitor. With a 180-pF capacitor, these transformers will tune to around 1.0 MHz. The 10.7-MHz transformers with a 180-pF capacitor will resonate at around 5.8 MHz. The regular 455-kHz i-f transformers with a 47-pF capacitor tune to about 700 kHz. Double-tuned transformers were made from 10.7-MHz discriminator assemblies, and center-tapped transformers to couple from the output of an MC1496 mixer to the LM373H were also made.

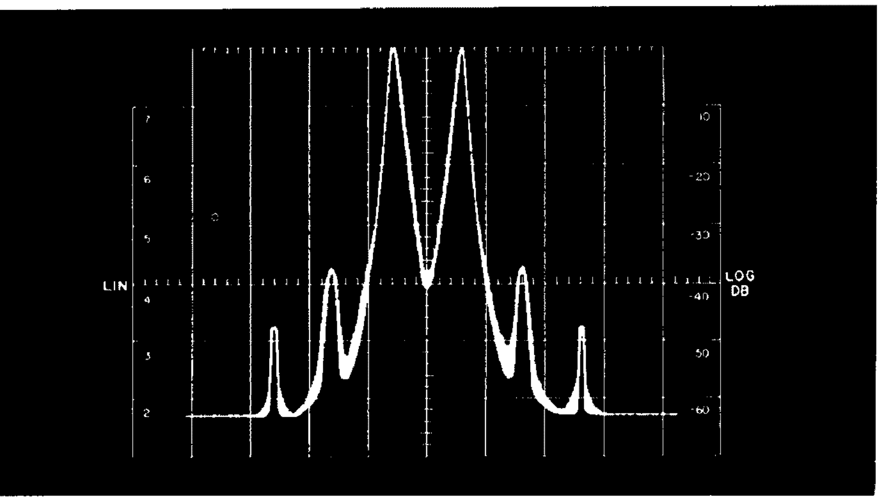
At the time this receiver was being developed, I was corresponding with Al Bernard, WA2JTN, who happened to have some very small knobs which he provided for this project. They are only 5/16-inch (79 mm) in diameter. They are ideal for this application but, unfortunately, they are in woefully short supply, so you'll have to settle for the smallest you can find.

As for the rest of the parts, I know you can't buy the Murata ceramic filters any place so I'll keep some on hand. For a complete list of parts I can supply, please send along a self-addressed, stamped envelope with your inquiry.

reference

1. Ray Megirian, K4DHC, "Using The LM373," *73 Magazine*, April, 1972, page 37.

ham radio



intermodulation-distortion

measurements on ssb transmitters

A comparison of IMD measurements of amateur ssb equipment shows that industry standards should be set by the EIA

Marv Gonsior, W6VFR, 418 El Adobe Place, Fullerton, California 92635

Specifications for the suppression of odd-order distortion products in amateur ssb transmitting equipment generally appear to be vague, inadequately defined and relegated to the sole domain of the equipment manufacturers. One observation might be that some amateur ssb specifications seem to be generated in sales departments rather than in engineering departments. Rarely, if ever, are the manufacturers' transmitter intermodulation distortion products clearly stated in easily understood terms or on any uniform basis. For example, the third-order IMD specifications I have reviewed are stated merely as −30 dB, an apparently good number, if valid, by the present state-of-the-amateur art. However, an IMD specification stated in this way is ambiguous because it does not give the basis, the power level at which the measurement was made, nor other pertinent data such as frequency, worst or best case measurement, etc.

It is significant to note that all intermodulation distortion power is valueless and only causes interference to adjacent channels of operation. The purpose of this article is to explore some of the foregoing with respect to the Collins S-Line and some other contemporary equipment using measurement data taken by relatively sophisticated laboratory test equipment which, operating under a quality system, has certified calibration accuracy traceable to the National Bureau of Standards.

I used the Hewlett-Packard Model 141S Spectrum Analyzer Display Section with the 8553L rf and 8552A i-f heads for making the measurements. This instrument is a recognized standard in the electronics industry with a specified calibration accuracy of a 0.5 dB. It covers

the range of 1 kHz to 110 MHz, with interchangeable i-f and rf heads and has variable selectivity. My tests were conducted at 100-Hz bandwidth which is necessary for accuracy and ease of measurement. The instrument's spurious responses, with a -40 dBm signal to the input mixer, are better than 70 dB down. All analyzers have spurious responses since they are all sensitive to overload, so it is especially important to observe this critical parameter—otherwise it will be evidenced by false, spurious presentations.

IMD measurements

The methods of making IMD measurements, together with considerable discussion have been covered in several excellent, pertinent publications.[1,2,3,4] These references reveal that IMD measurements may be taken on one of two bases. In the conservative method which I used, the spurious level is referenced to *either* tone of a two-tone test. The other method is also acceptable, but is a less conservative method; it compares one distortion product to the sum of two tones, or the PEP. This yields a more impressive number, resulting in a -6 dB *apparent* improvement. This is because of

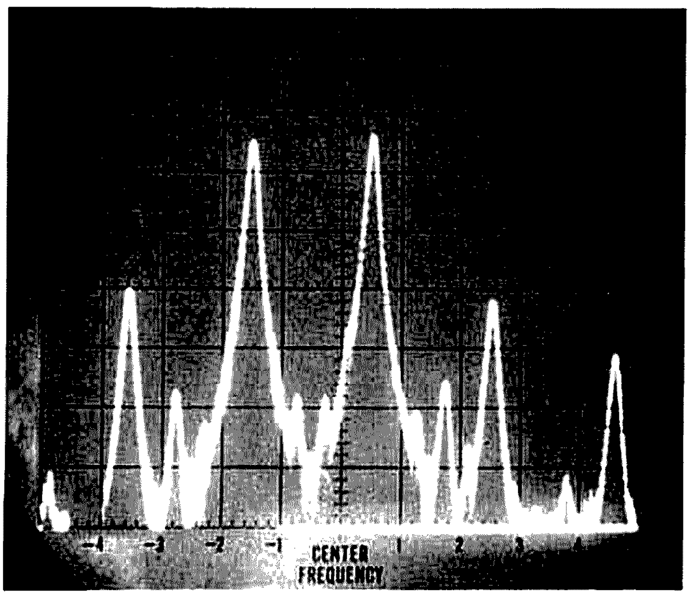


fig. 1. Spectrum analyzer display of intermodulation distortion test on a Signal One CX7A. Two-tone test, 800- and 1800-Hz tones. Horizontal scale is 500 Hz per division. Third-order products are 25-dB down (see text).

the fact that the PEP value, or the sum of the two tones, is twice the voltage, providing a 6-dB greater reference level. Therefore, with that power level as the reference, the distortion products would *appear* to be 6-dB further down. A third method that is widely used in military work is the Intercept Point method[5] which resolves most of the ambiguities. However, I will not go into that here.

Most vacuum-tube manufacturers, and military and commercial suppliers tend to specify the distortion products of their equipment using the conservative method. Quite naturally, equipment manufacturers may indulge in some "specsmanship" by publishing performance specifications which tend to favor themselves. During the preparation of this article I contacted the National Bureau of Standards, the IEEE and the EIA to ascertain whether any standards had been published on the subject of ssb transmitter IMD measurement. They indicated that none were published. A proposed EIA specification was prepared but, unfortunately, it has never been published; it was to be based on the liberal basis; i.e., the PEP reference. I have therefore concluded that the amateur equipment manufacturers in the United States, with the one notable exception of Collins, have chosen to use the liberal interpretation while the rest of the industry and the military appear to adhere to the conservative method. The point is that amateurs should be aware of the differences and the implications of both.

For purposes of clarity in this article, I will refer to the two bases as the conservative and liberal methods and, unless stated, all references of mine will be on the conservative basis. In other words, you should add -6 dB to all my data to make it consistant with most of the published specifications and test data on amateur ssb equipment. For example, take the RCA tube specification sheets for the very popular 6146B type tubes. The manufacturer states that these tubes exhibit -24 dB CCS and -26 dB ICAS for their third-order products at PEP power output levels of 49 and 61

watts, respectively, "referenced to either of the two tones and without the use of feedback to enhance linearity," i.e., the conservative method.

Eimac makes the more explicit statement: "The intermodulation distortion products will be as specified or better for

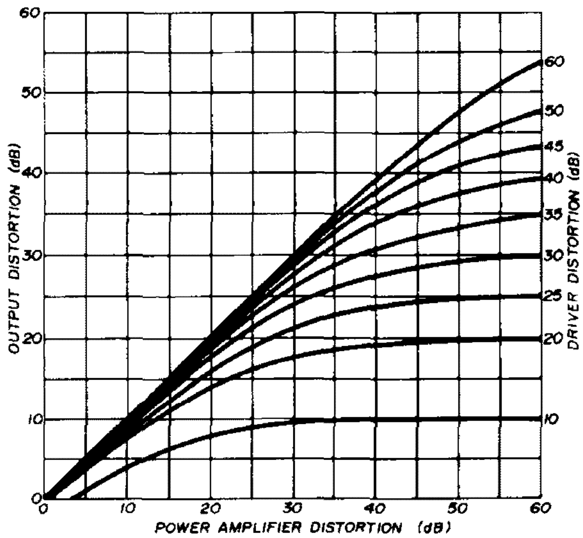


fig. 2. How IMD levels of the driver and output stage combine to determine overall IMD performance (courtesy Collins Radio Company).

all levels from zero-signal to maximum output power and are referenced against one tone of a two-equal-tone signal." That statement is extremely significant, as we shall see. Allowing for tube variations in performance, matching, aging and power supply regulation, combined with other miscellaneous factors such as neutralization, drive regulation, bias, etc., it is impossible to achieve a conservative −30 dB third-order IMD at the 100-watt level from the 6146 family of tubes without the use of negative rf feedback. Bill Orr showed that sweep tubes generally exhibit less than this performance when operated near or at full power, e.g., in the −18 to −22 dB region for the third-order products.[1]

Looking at the data sheets for the 8072, the output tube used in the Signal/One CX7A, RCA specifies, on the conservative basis, −30 dB on the third-order products at 80 watts PEP power output. Signal/One specified their equipment as a "nominal 300-watt PEP input and 160-watt PEP output." They further specified that their third-order IMD was "greater than 30 dB below each of two equal tones at full rated output." This specification is on the conservative basis and, extrapolating from the tube data sheets, it is an obvious impossibility because of the tube manufacturer's 80-watt PEP output specification versus the equipment manufacturer's 160-watt PEP output specification, but it surely makes the meters dance. For my IM test results of −25 dB at full power out on a CX7A which had just been overhauled by the factory, see **fig. 1**. The exciter did, however, meet its −30 dB IMD specification at 100 watts PEP output.

tube operating conditions

During the course of my measurements, a friend brought in his Yaesu FTDX560 for a quick check. The third-order products, at full power out, predictably were −22 dB. At 40 mA of quiescent plate current, 10 mA less than that specified, the third-order products measured −20 dB. It was interesting to observe that the fifth-order products were measured at −50 dB at 60 mA of quiescent plate current and were only −35 dB at 40 mA. Throughout my measurements, this type of result was common. Since the third-order products are generally the

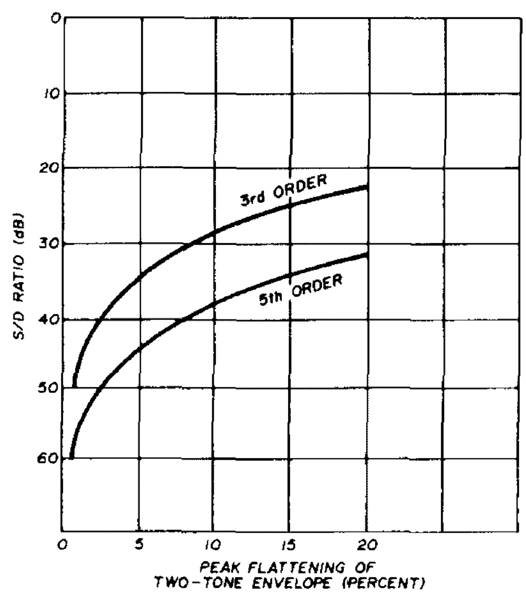


fig. 3. Relationship of two-frequency envelope peak flattening to signal-to-distortion ratio.

worst offenders, I have not commented on the fifths. However, in a local or short skip situation, *all* products play an important role.

One significant fact I found was that the third-order products could be improved 6 dB by merely resetting the quiescent plate current up to 60 mA from the specified 50-mA level. A note of caution is in order here, since this probably would affect tube life under full-bore operation because of the additional heat, and TV sweep tubes are very dissipation limited. It is important to note that most amateur equipment is operated at or near the full capability of the output tubes.

Looking at the practical situation, with some signal degradation from linear amplifiers, most amateur stations are probably transmitting signals with third-order IMD products in the −20 to −24 dB region. By way of contrast, commercial and military equipment is generally specified in the −30 to −35 dB range. State-of-the-art military equipment is specified in the −40 dB range with −50 dB attainable. It is clear that IMD is less in linear stages than in non-linear stages and becomes progressively worse as the linearity of a stage is compromised from class A to B1 and B2 operation. On the other hand, the efficiency of a power amplifier is significantly higher for less linear operation. This means that the lower power stages may be operated inefficiently at class-A bias to provide very low IMD levels while the output stage is usually biased for higher efficiency and increased IMD levels. If output stage efficiency is a driving factor, the output stage IMD performance will limit the overall performance of the equipment. The way in which the IMD levels of the driver and the output stage combine to determine the overall performance is illustrated in fig. 2.*

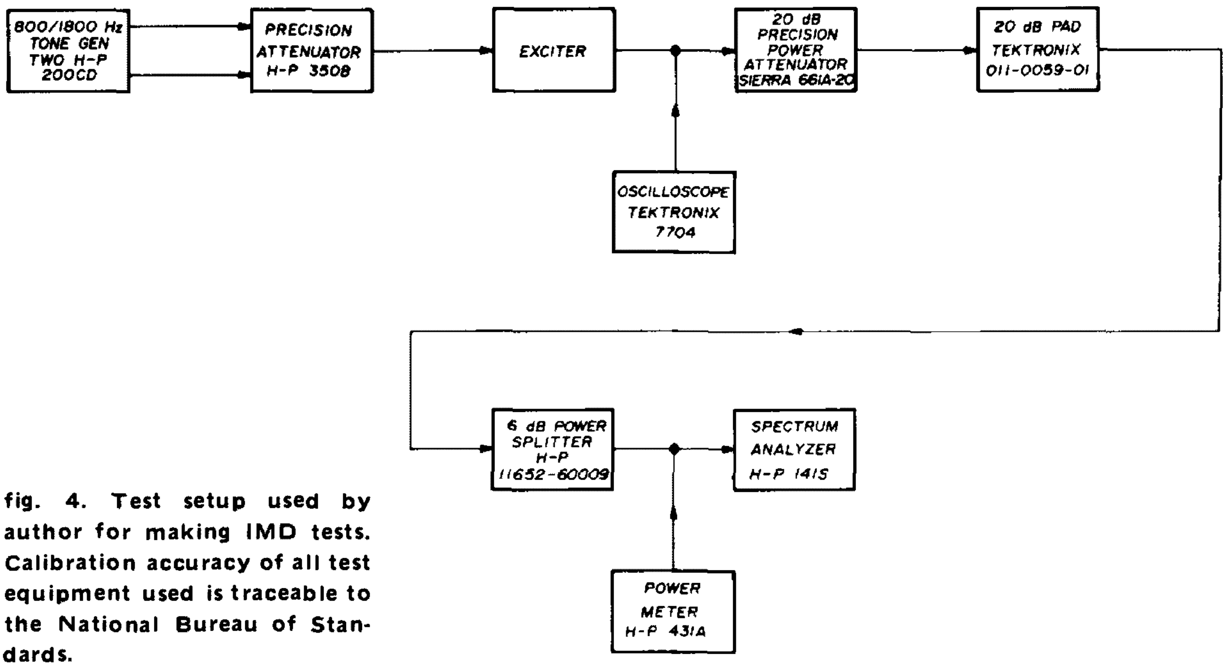


fig. 4. Test setup used by author for making IMD tests. Calibration accuracy of all test equipment used is traceable to the National Bureau of Standards.

In a period of time when we are concerned with already crowded amateur bands, IMD specifications are worthy of consideration. I do not presume to recommend military specs for amateur equipment, but I do urge amateurs to strongly consider the credibility of the published specs, look at the tube data sheets, and operate their equipment in a reasonable manner. By this I mean something less than maximum power, using an

*Distortion measured in dB below one tone of a two-tone signal. Products are assumed to add in phase. Fig. 2 is derived from a plot of the relation

$$x = 20 \log 10^{y/20} \quad 1$$

where x is driver distortion minus amplifier distortion, and y is driver distortion minus output distortion.

adequate monitoring and control system combined with all the other facets of the overall problem. Now you may understand why some rigs sound and look better than others.[6]

I wrote letters to each of the five major U.S. amateur ssb equipment manufacturers requesting data on their IMD specifications and measurements. Three of the four who replied stated that they use the PEP reference as I indicated earlier. The two largest Japanese manufacturers were not quiried, as they do not specify IMD. Since they use sweep tubes or 6146s at full power without rf feedback, their third-order IMD products can readily be predicted to be between −18 and −26 dB; with sweep tubes, the −18 to −22 dB region.

Collins S-line

Let's take a look at the IMD performance of Collins' equipment which was designed over fifteen years ago, but which probably exceeds the IMD suppression performance of all current production amateur communications equipment. Two features make this possible. One is an excellent two-stage alc system combined with the use of rf negative feedback. The advantage of this is simply that a rig with 10-dB of rf negative feedback will be essentially 10-dB cleaner than a rig without it. To this date, no other production amateur equipment in the world is known to use rf negative feedback. As I previously pointed out, RCA specifies the 6146As and 6146Bs at −26 dB, to which −10 dB of feedback at 14 MHz is added, so Collins' −30 dB spec can be easily met or exceeded, allowing for all variations.

It is significant to note that every piece of amateur ssb equipment, both linears and exciters, manufactured by Collins employs rf negative feedback! Collins advises that *every* S-Line, KWM-2, and linear amplifier goes through an IMD measurement test in their production acceptance test procedure. That unique type of quality control for amateur equipment is to be commended. Collins also states that tubes are the major reason for rejection for IMD products so this would indicate that the tubes should be given much more consideration than is normally the case.

For example, in my Collins 32S-1 exciter I always match the 6146s for equal quiescent plate current. It is quite revealing to note the wide variation of this single parameter from the same manufacturer. In the course of making my measurements, it was noted that the very smallest amounts of indicated grid current on the 6146s resulted in a substantial increase in the distortion levels. This indicates that any departure from class AB1 is totally unwarranted and should serve as a warning to those who would defeat an alc system in any fashion. Pappenfus discusses the resulting flattopping in two chapters of his book (see fig. 3).

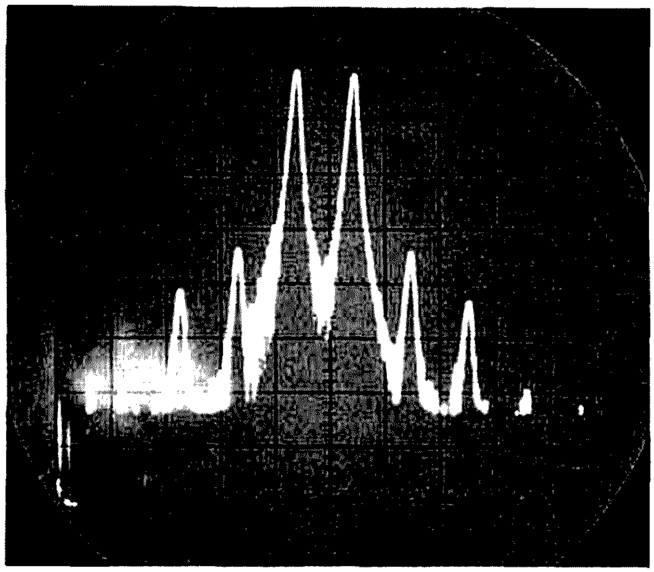


fig. 5. IMD distortion of Collins 32S-1 exciter at the 100-watt PEP level. Third-order products are about 35-dB down; fifth order products are more than 40-dB down.

Now let's look at some measurements taken on my 32S-1 exciter, which is more than thirteen years old. Third-order product measurements were made in an engineering calibration laboratory using the block diagram shown in fig. 4. The result at the 100-watt PEP power output level is shown in fig. 5. Sutherland and Orr have shown that the third-order products are generated on a kinked curve

due to distortion cancellation,[2] so my power level tests were considered appropriate to preclude errors. Also, it was desirable to see if tube balance by plate idling current is significant and if there was any real difference between 6146As versus 6146Bs. RCA states there is none. My test data tends to confirm this, but my sample size was very small. Further, the effects of varying the various operating parameters of the 6146s are shown in fig. 6.

During the course of taking this data, despite using two husky fans, it was noted that heat dissipation is a major constraint. Repeatability of the test data was heavily dependent upon proper cooling and reasonable operating periods to minimize the thermal impact. Therefore, serious consideration should be given to this parameter during normal operating, and especially under contest conditions. One first thought is to open the cabinet cover. Next, of course, would be a small, quiet fan mounted off the chassis to preclude microphonics. It must be recognized that the two-tone test is a very strenuous exercise of the equipment since the average power is substantially higher than with speech.

I thought it would be valuable to ascertain the IMD level of the 6CL6 driver stage in the 32S-1 to determine if it could be improved, thereby improving the 6146B output stage IMD. It measured a startling −22 to −29 dB on the third-order products, depending on the power level, a figure that at first was of very serious concern. Later it was determined that, with the feedback loop closed, what I was reading was the net IMD after subtracting the counter distortion being used to linearize the output tubes. Without any doubt, the driver, which is essentially operating class A, is producing IMD in the −40 dB region. No measurable improvement in the driver IMD was attained, as read directly or indirectly in the power-amplifier linearity, with any change in its operating biases.

I would like to describe some of the modifications which were installed in my 32S-1 exciter to increase its linearity. Capacitor C61, a 10-pF ceramic which forms the basic feedback network in the S-Line, has been placed in parallel with a Corning glass 5.6-pF capacitor. This provides a measured third-order improvement of 3 dB at 14 MHz at full power output. Another simple mod, although not affecting the IMD, provided a de-

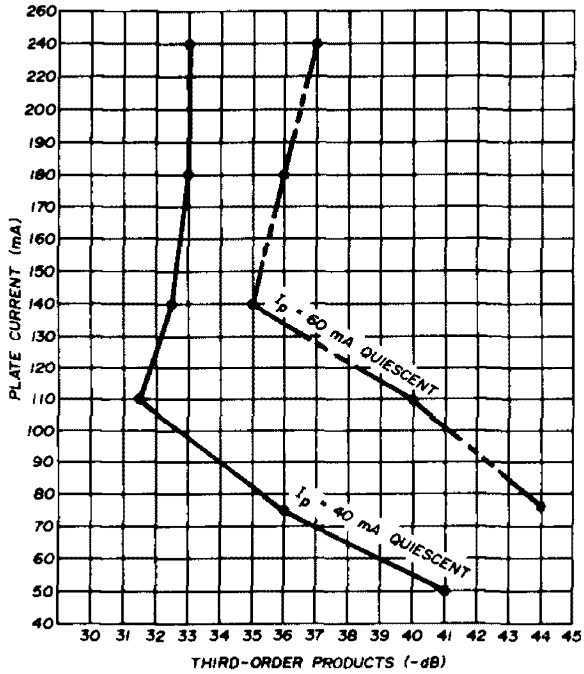


fig. 6. Effect of varying Collins 32S-1 quiescent (idling) plate current on third-order distortion products. The 40-mA level is normally used but 60 mA quiescent plate current improves IMD. During these tests plate voltage was 820 volts and screen voltage was 275 volts.

crease in the audio harmonic distortion—this is the addition of a 250k resistor between the plates of V1A and V1B, the audio amplifiers.[7] This results in an audible improvement in the speech quality. I note that the same principle is applied in the Yaesu FTDX560.

To make up for the loss of gain due to the inverse feedback, the vox amplifier, V15A, will require a 10-μF bypass from the cathode (pin 3) to ground. Otherwise the vox will probably cease to operate as there is about 15-dB loss of audio signal, the amount of the feedback, through the system. Further, I set the 6146B quiescent plate current at 60 versus the 40 mA which is specified by Collins. This yields an additional 4-dB improvement in the

third-order distortion products. Of course, this might cause a heating problem and therefore affect tube life in some equipment, but this is not so in my case as I require only 40 watts of PEP drive for my class AB1 linear amplifier.

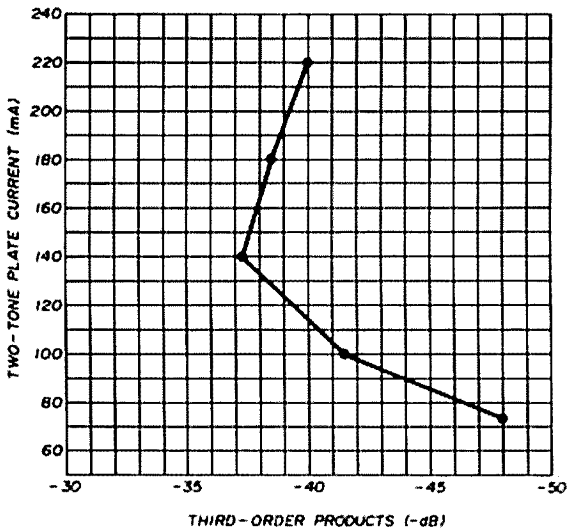


fig. 7. Third-order IMD performance of Collins 32S-1 exciter modified by the author (see text). Test conducted at 14.25 MHz; plate voltage, 780 Vdc; screen voltage, 240 Vdc; quiescent (idling) plate current, 60 mA.

Also, I run the 6146B screens from the basic low-voltage supply rather than the bleeder arrangement in the original design.[8] This improved screen voltage regulation will increase the power output and also should slightly improve the IMD, although no comparative measurement was made.

Tests were also made to determine the effects on IMD which would result from rf clipping. Only 1-dB of degradation was measured, at about 15-dB of clipping. Some of this is probably due to the fact that the alc system is being partially deceived through the loss of the high peaks by the clipping, as well as the natural result of the higher average power level. The combination of these modifications allowed a reduction of IMD at the 100-watt output level from −32 to −39 dB, as shown earlier. Fig. 7 shows the results produced by reducing the plate and screen voltages in an effort to find an optimum operating condition. The power output was down to 60 watts PEP but the desired IMD reduction was attained.

On the subject of tubes, there are stories circulating that 6146As are cleaner than 6146Bs, etc. I wanted to determine just what advantages, if any, there might be between the two types. A series of tests was conducted using matched pairs for quiescent plate current, unmatched tubes, and mixed 6146As and 6146Bs. The results were relatively inconclusive with up to 3-dB variations measured. My sample size was too small to develop any significant trends, so I would stick to the published literature, i.e., use well-matched tubes that are in good condition and run them cool at less than maximum power, neutralize carefully, load properly, etc.

While the test setup was available, I decided to measure the harmonic suppression of my S-Line. It measured −47 dB on the second harmonic and greater than −60 dB on the fourth while operating at 14.25 MHz into a matched 50-ohm load.

other equipment

My basic data are concerned with the Collins S-Line, but other measurements on a limited basis are also presented to

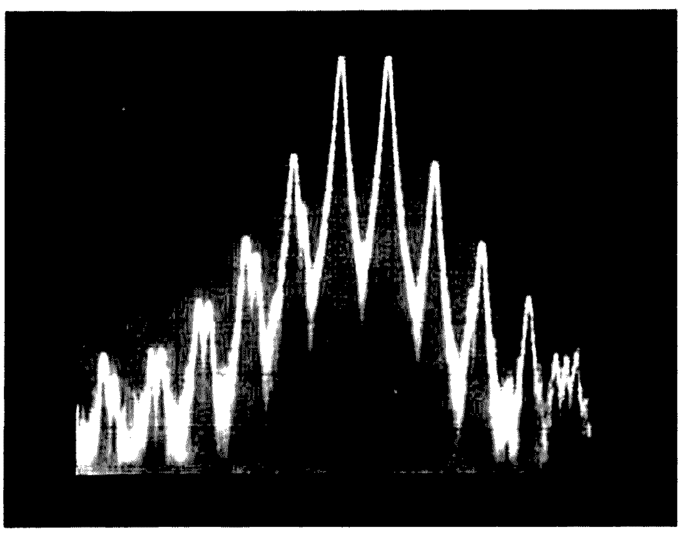


fig. 8. Spectrum analyzer display of two-tone test on Kenwood T599 ssb transmitter, referenced to one of two equal tones. Maximum two-tone PEP output without limiting was 72 watts as observed on Tektronix 585A oscilloscope in conjunction with a Hewlett-Packard 431B power meter. This display shows third-order products down −19 dB, fifth-order products down −34 dB.

support some of my earlier statements. From these it may be seen that, with the exception of Collins, the conservative, -30 dB IMD specifications are not attainable at full power levels, regardless of the specifications published by the manufac-

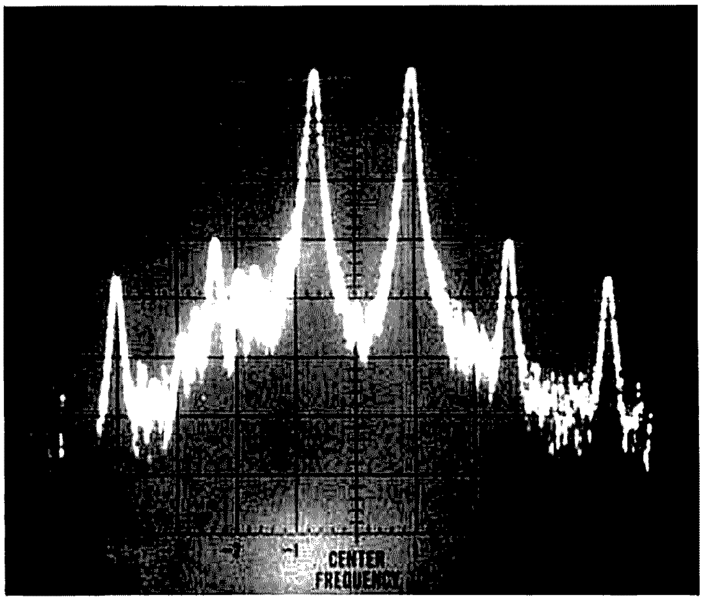


fig. 9. Spectrum analyzer display of two-tone test on Collins KWM-2A ssb transmitter, referenced to one of two equal tones. Maximum two-tone PEP output without limiting was 100 watts. This display shows third-order IMD products down -30 dB, fifth-order products down -37 dB.

turer. In one case, the Kenwood T599, the IMD specs are not published by the manufacturer. My measurements at 72 watts PEP output of -19 dB on the third-order products differ with those published[9] by 8 dB as shown in fig. 8. Two other amateurs, using the same measurement equipment, confirm my data.

Fig. 9 shows the measured -30 dB results taken on an old Collins KWM-2 at the 100-watt PEP output level. It was sorely in need of total refurbishment, but it still meets the published specifications.

conclusion

If you are a concerned amateur, and we all should be, let me say, *caveat emptor*! The IMD performance may be more or less readily determined to within a couple of dB by examining the output tubes, considering their specified input power, and reading the literature. What I have found has been clearly supported in all the text books. If we want to radiate cleaner signals, there are not any magic formulas to accomplish that. I would, however, lobby for one published standard method of IMD measurement so that everyone understands the related specification and measurements. Only one manufacturer of amateur equipment, Collins, specifies their IMD suppression on the conservative basis, although even they do not make it clear in their literature. All other amateur manufacturers adhere to the liberal basis which provides them with a relative 6-dB advantage. It is a valid measurement, though somewhat misleading, and is often misunderstood and certainly misquoted. It now becomes obvious that in order to compare the proverbial oranges, all manufacturers should use one specification standard.

I would like to express my sincere appreciation to Bob Bruemmer, WB6CGG, who spent many hours setting up the measurement equipment used for taking the data presented and verifying its calibration.

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ham radio

modern rf amplifiers for communications receivers

A discussion of the performance characteristics of modern rf amplifier stages, and how to choose the circuit best suited for your own application

Ray Moore, 116 Walnut Street, Walpole, Massachusetts 02081

Since James Lamb first described the modern communications receiver in 1932 amateurs have expected a good receiver to have one or two rf amplifiers. There was good reason for this in the 1930s as mixers were noisy and even the relatively noisy rf amplifiers which were available were an improvement. The rf amplifiers were also necessary to compensate for the losses in the tuned circuits required to attain a decent image ratio with single conversion to an i-f around 455 kHz.

The pentode tube was used almost exclusively in rf amplifiers for thirty years. Many amateurs will remember the succession of tube types—58, 6D6, 6K7, 6SK7, 6AK5, 6BA6, 6DC6, 6BZ6 and 6EH7. Bipolar transistors were used briefly during the 1960s, not because they made better rf amplifiers, but to satisfy the demand for all solid-state receivers. The junction fet (jfet) and dual-gate mosfet soon appeared and replaced the bipolar devices in new designs.

An rf amplifier does several things for a receiver. First of all, it provides gain to make up for any losses in the rf tuned circuits. Secondly, it provides low- noise gain to override mixer noise. It can also provide an extra gain-controlled stage and can be used to perform impedance-matching functions.

However, the rf amplifier also has disadvantages which must be weighed against the advantages in any particular application. Most important, since it places an additional unprotected stage ahead of the selective filters, it can aggravate mixer nonlinearity by amplifying close-by, unwanted signals.

antenna noise figure

The receiving system is a chain consisting of the antenna noise figure, the input stage, the second stage and so on. If the chain is properly designed the overall system noise figure is determined by the first link in the chain, the antenna noise figure.

The ultimate factor in deciding what weak signals you can hear is the signal-to-noise ratio between the signal and the atmospheric, galactic and manmade noise impinging on your antenna. If the signal is equal to the noise level you can at least detect it and possibly carry on marginal communications. Therefore, the receiver

should be quiet enough for you to hear the ambient noise level. Anything much quieter than that is superfluous.

The atmospheric noise level has been measured and is available in terms of antenna noise figure.[1] **Fig.** 1 shows the antenna noise figure for the northeastern United States between 0800-1200 local time in the winter. The noise falls within the shaded area 80% of the time, above it 10% of the time, and below it 10%. In addition, there is galactic noise striking the earth's atmosphere at all times. The galactic noise level is shown by the dashed curve for all frequencies but it can only penetrate to your antenna at frequencies above the critical frequency.

A third curve extending up to 10 MHz gives the average level of manmade noise at a quiet receiving location. It is often the limiting factor at frequencies below 10 MHz.

What is the lowest antenna noise figure you can expect in the high-frequency range? The figure of 18 dB is often quoted as the limit and fig. 1 shows that this is the galactic noise level at 30 MHz. The critical frequency seldom reaches 21 MHz so galactic noise is still the determining factor on that band.

On 20 meters the critical frequency may sometimes be high enough to shut out galactic noise but the atmospheric noise is above 28 dB 90% of the time. On 40 and 80 meters manmade noise puts a floor under the minimum antenna noise figure. Therefore, the assumption that the lowest antenna noise figure encountered on hf is 18 dB is theoretically valid.

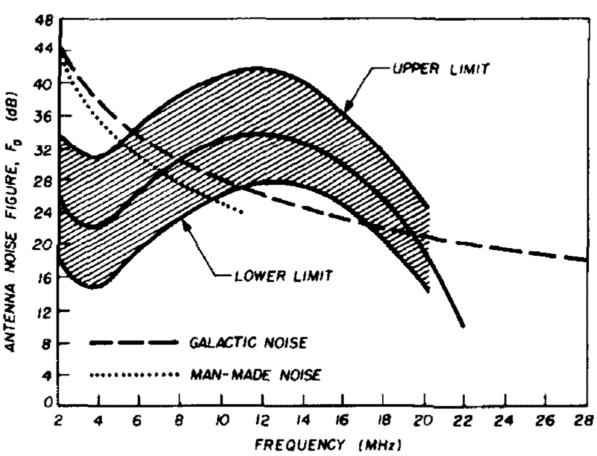


fig. 1. Atmospheric noise levels expressed as antenna noise figure for 0800 to 1200 local time on a typical winter day in the northeastern United States.

There is reason to believe, however, that on some few occasions the noise level on 10, 15 and even 20 meters drops below the predicted values, perhaps down to 10 dB or less.[2,3] Little information is available to explain these occasional quiet periods. Perhaps they are due partly to unusual absorption of galactic noise as it passes through the atmosphere. As you go *higher* in frequency, of course, the galactic noise decreases so that the antenna noise figure is 13 dB at 50 MHz and 6 dB at 100 MHz.

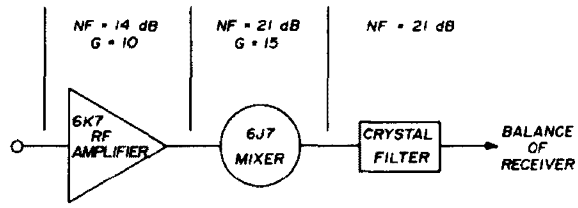


fig. 2. Noise characteristics of a typical communications receiver of the 1930s.

The overall noise factor of a system composed of a receiver and an antenna is

$$F_{sys} = F_a - 1 + F_r \qquad (1)$$

where

F_{sys} = overall system noise factor (ratio)
F_a = antenna noise factor (ratio)
F_r = receiver noise factor (ratio)

Assume you are operating a receiver with a 10-dB noise figure (noise factor = 10) on 10 meters where the antenna noise figure is 18 dB (noise factor = 63). How much does the receiver degrade the overall noise figure?

$$F_{sys} = 63 - 1 + 10 = 72 \text{ or } 18.6 \text{ dB}$$

For a perfect receiver (zero-dB noise figure) the calculation is

$$F_{sys} = 63 - 1 + 1 = 63 \text{ or } 18 \text{ dB}$$

Thus, the receiver degrades the noise figure of the system by only 0.6 dB.

The story is different at vhf. Suppose

you are operating at 100 MHz with the same receiver. Here the antenna noise figure is only 6 dB (noise factor = **4**).

$$F_{sys} = 4 - 1 + 10 = 13 \text{ or } 11.1 \text{ dB}$$

In this case the NF is degraded by 5.1 dB over that of a perfect receiver.

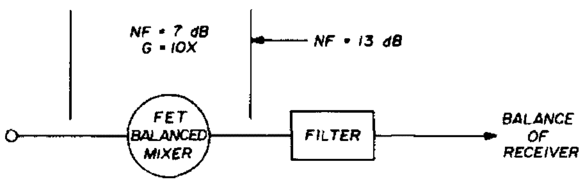


fig. 3. Noise characteristics of a modern fet balanced mixer feeding an i-f strip.

receiver noise figure

One rule-of-thumb says that the first stage of a well designed receiver should contribute all except one dB of the overall receiver noise factor. The overall noise factor of a receiver can be found from

$$F_r = F1 + \frac{F2 - 1}{G1} + \frac{F3 - 1}{G1G2} \qquad \textbf{(2)}$$

where F_r = overall receiver noise factor (ratio)
F1,F2,F3 = noise factors (ratio) of first, second and third stages
G1,G2 = power gain (ratio) of first and second stages

Apply this formula to the receiver shown in fig. **2** which is typical of receivers used in the late 1930s.

$$F_r = 25 + \frac{126 - 1}{10} + \frac{126 - 1}{10 \times 15}$$

$$= 25 + 12.5 + .83$$

$$= 38.3 \text{ or } 15.8 \text{ dB}$$

Notice how little the second and succeeding stages contribute to the overall noise factor.

Let's look at another example. **Fig. 3** shows a modern fet balanced mixer working from the antenna into an i-f strip with a NF of 13 dB (noise factor = 20). The mixer, including its input circuits, has a NF of 7 dB (noise factor = 5) and a gain of 10 dB.

$$F_r = 5 + \frac{20 - 1}{10} = 6.9 \text{ or } 8.4 \text{ dB}$$

Substitute a modern diode balanced mixer with a 6-dB NF (noise factor = 4) and 0.4 gain for the fet mixer in fig. **3** and you obtain

$$F_r = 4 + \frac{20 - 1}{0.4} = 51 \text{ or } 17.1 \text{ dB}$$

This receiver obviously needs an rf amplifier to reach an acceptable operating noise factor.

rf amplifier noise figure

A simplified schematic and equivalent circuit of an rf amplifier is shown in fig. **4**. The noise factor of such a stage below 20 MHz is approximated by

$$F = 1 + \frac{R_A'}{R_D} + \frac{R_{eq}}{R_A'}\left(1 + \frac{R_A'}{R_D}\right)^2 \qquad \textbf{(3)}$$

where F = amplifier noise factor (ratio)
R_A' = transformed antenna resistance = $n^2 RA$
R_D = dynamic resistance of tuned circuit = QX_L
R_{eq} = equivalent noise resistance of amplifying device (fet or tube)

The first two terms, $1 + R_A'/R_D$, give the noise contribution of the input tuned circuit. When the antenna is exactly matched to the receiver $R_A' = R_D$ and the noise factor of the input tuned circuit is 2, or 3 dB.

The third term, $(R_{eq}/R_A')(1 + R_A'/R_D)^2$, is the noise contribution of the fet or tube and reduces to $4R_{eq}/R_A'$ when the antenna is matched to the receiver. Above about 20 MHz the input impedance of both tubes and fets falls off, shunting the dynamic resistance, R_D, adding additional factors to the calculations.

When the antenna is matched to the receiver the NF of the input circuit is 3 dB. This NF can be made to approach zero dB by reducing the transformed

antenna resistance, R_A'. Overcoupling by increasing the primary turns, which *reduces* the ratio n^2, will decrease R_A'. This technique is sometimes useful at vhf when the noise factor of the matched amplifier is less than 6 or 8 dB. However, this is done at the sacrifice of rf selectivity since R_A' effectively shunts R_D, lowering the Q of the tuned circuit. The optimum value of R_A' which gives the lowest possible NF is

$$R_A' \text{ (optimum)} = R_D \sqrt{\frac{R_{eq}}{R_D + R_{eq}}} \qquad (4)$$

input tuned circuits

The input tuned circuit(s) of an rf amplifier has two functions. It matches the antenna to the active device and it provides rf selectivity. Fig. 5 shows various ways the single tuned circuit may be used as an input circuit.

The most common input circuit is the untuned primary to resonant tank circuit shown in fig. 5A. The size and coupling of the primary is determined by the characteristics of the antenna. Most modern receivers are designed for 50- or 75-ohm transmission lines and the primary is adjusted so that the reflected impedance, combined with the impedance of the primary itself, is the proper value. The match is rarely exact and will vary from one end of a band to the other.

Receivers with low-impedance inputs are sensitive to the characteristics of the antenna and line to which they are connected because there is a large step-up ratio from primary to secondary. Older receivers used 200- to 400-ohm primaries to match the open transmission lines of the day and were more tolerant of the type of antenna used.

The tapped coil method of coupling the antenna is often used at vhf (fig. 5B) because it allows convenient adjustment of the coupling to produce optimum r_A' for minimum NF. Using a noise generator to determine NF, a fine wire is successively soldered up from the bottom of the coil until the best NF is found. The output impedance of the coil can also be varied by tapping down, as shown.

The classic R390 military receiver has a versative input circuit that will handle random-length wires, balanced or unbalanced lines (**fig. 5C**). A random-length wire or whip is connected to the top of

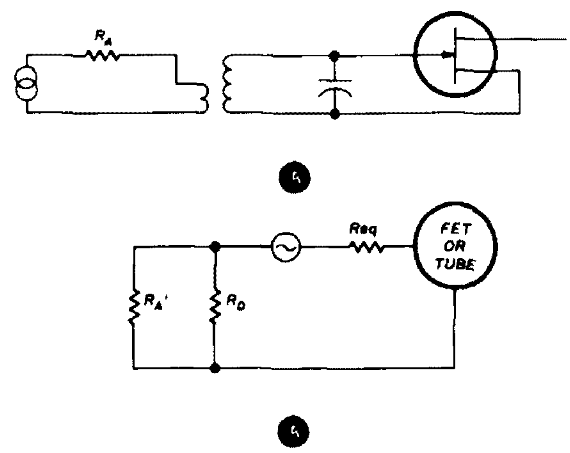


fig. 4. Noise sources in the antenna input stage of a communications receiver. Simplified circuit is shown in (A). Equivalent circuit is shown in (B).

the tuned circuit through C1 (input 1). The value of C1 varies from 5pF on the 16- to 32-MHz range to 120 pF on the 0.5- to 1-MHz band. Balanced lines are connected to the outside terminals of input 2 and are electrostatically balanced to ground by the capacitive divider. Unbalanced transmission lines are connected to one side of input 2.

One method of obtaining a precise impedance match at a particular frequency is shown in **fig. 5D**. This circuit became prominent when it was used in the famous R9'er impedance-matching preamplifier in the late 1940s.[4,5] The coil and variable capacitor are adjusted successively for maximum output until no further improvement is obtained. The circuit was said to match resistance from 58 to 3120 ohms when the coils were loaded with a 7k swamping resistor and the variable capacitors had a range of 5 to 100 pF.

The pi network allows precise adjustment of both the input and output impedances (**fig. 5E**). It is sometimes used

at vhf to adjust the coupling for minimum noise factor.

filters

The trend today is toward more complex circuitry between the antenna and the rf amplifying device. The ultimate rf selectivity is provided by crystal lattice filters. Conklin has described practical filters for 20 meters with stop bands 100-dB down and 6/60 dB shape factors better than 1.8:1.[6] He used a switch selected set of eleven filters, each 33-kHz wide, to cover the entire 20 meter band (fig. 6A).

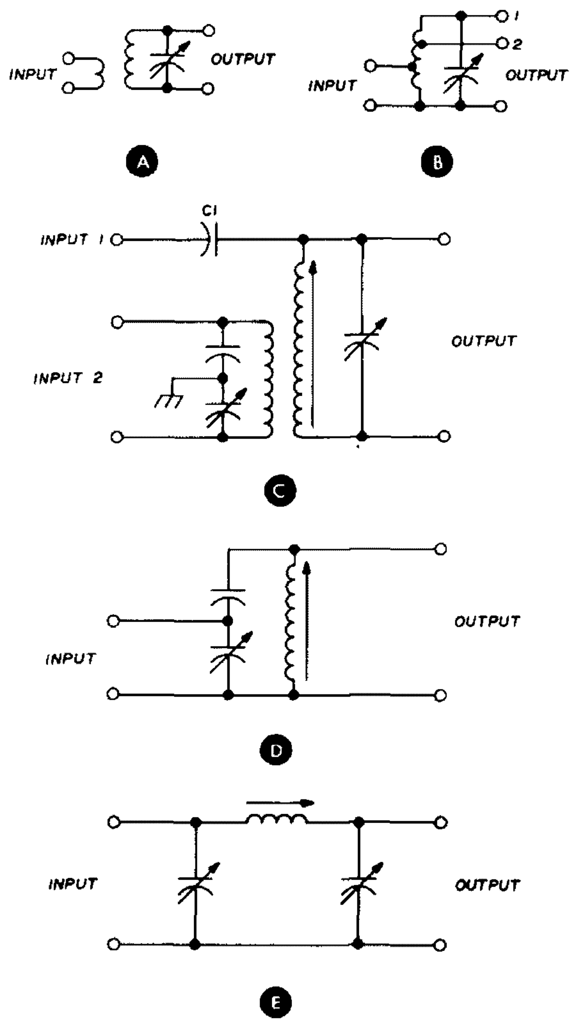


fig. 5. Antenna input circuits used in communications receivers.

Helical resonators are capable of Qs of 700 at the high frequencies and more than 1000 at vhf and several of them can be cascaded. A recent article[7] described practical helical resonators for 15, 20 and 40 meters (fig. 6B). Their disadvantages include size and their requirement for fixed tuning. A three-section, 15-meter filter measured 10½-inches (26.7-cm) long, 5¼-inches (13.3-cm) high and 3½-inches (8.9-cm) deep.

Multiple conventional tuned circuits can be mutual-inductance coupled, common- inductance or -capacitance coupled or top-capacitance coupled. The common-inductance coupled circuit in fig. 6C was used in a recent home-built 75-meter DX receiver.[8] The value of the coupling coil for that band was given as 0.3 μH.

A similar circuit using top-capacitance coupling is shown in fig. 6D. In one recent high-frequency receiver design C_c was a 0.2- to 3-pF ceramic capacitor.[9] The coupling was optimized by successively reducing C_c from maximum capacity in small steps and then returning the main tuning capacitors until there was a slight loss in gain as indicated by the S-meter.

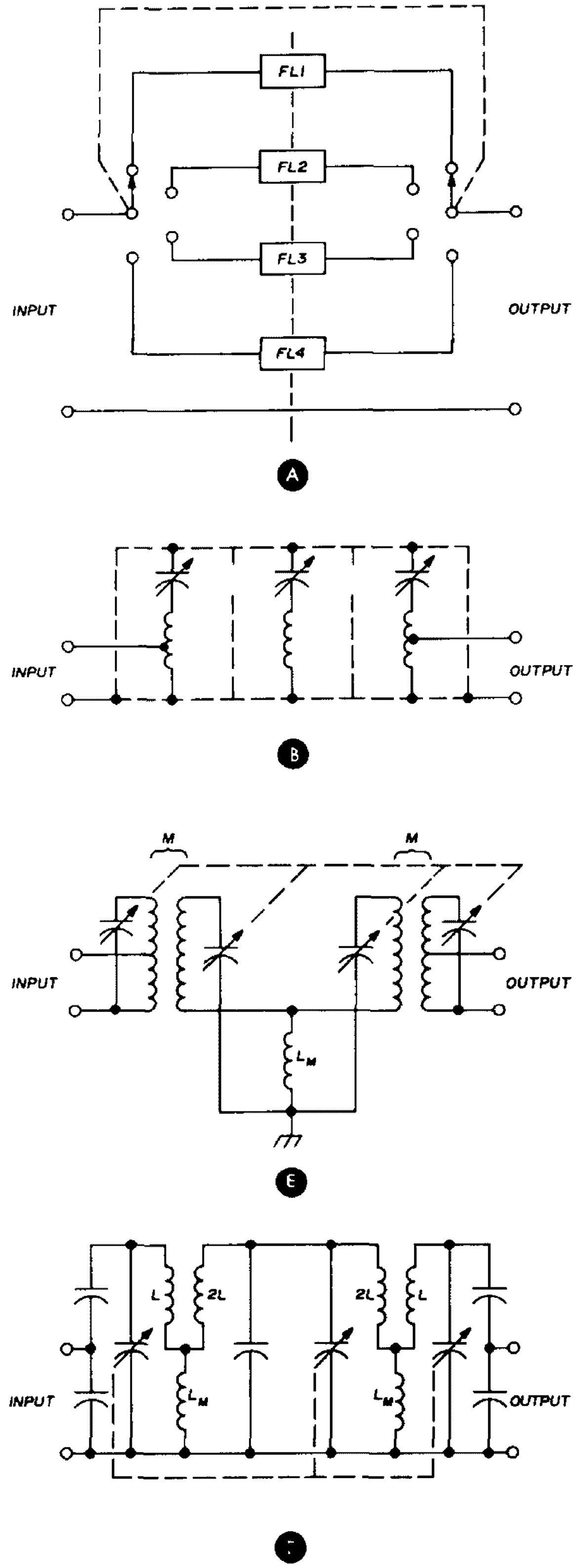
Old TRF receivers sometimes lumped their tuned circuits at the front of the receiver. One such preselector circuit used four tuned circuits with both mutual-inductance coupling and common-inductance coupling (fig. 6E).

Cohn has shown how to maximize stop-band attenuation while minimizing insertion loss for multicircuit rf filters.[10] One filter designed from his work is shown in fig. 6F.

selection of device and operating conditions

The rf amplifier designer has three devices from which to choose: vacuum tubes, bipolar transistors and fets. The vacuum tube can be eliminated because it is not compatible with the modern solid-state receiver in terms of heat generation, size, reliability and power supply requirements. The bipolar transistor is deficient in signal-handling ability. This leaves the fet as the almost universal choice for high-frequency rf amplifier applications.

When selecting an rf amplifier device the designer must consider such char-



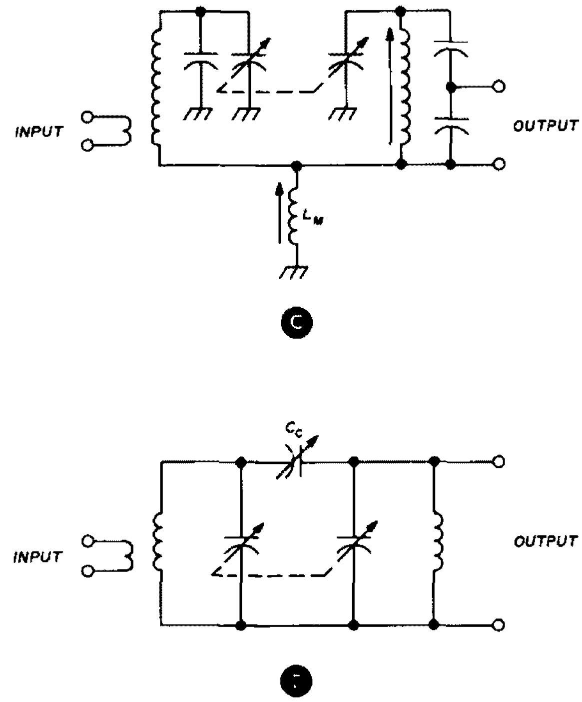


fig. 6. Multi-element rf input filters used in communications receivers. Use of each is described in the text.

acteristics as gain, noise figure, linearity, feedback capacitance (stability) and adaptability to manual and automatic gain control. The relative importance of each of these characteristics varies from one application to another. Table 1 lists the characteristics of some of the more popular rf fets.

Noise and gain are important design considerations in the upper vhf region but are secondary to linearity at the high frequencies. An idea of the noise performance of fets is given by the equivalent noise resistance formulas.

$$R_{eq}(\text{jfet}) = \frac{1}{G_m} \tag{5}$$

$$R_{eq}(\text{mosfet}) = \frac{0.67}{G_m} \tag{6}$$

$$R_{eq}(\text{triode}) = \frac{2.5}{G_m} \tag{7}$$

$$R_{eq}(\text{pentode}) = \frac{I_b}{I_b + I_{c2}}\left(\frac{2.5}{G_m} + \frac{20\, I_{c2}}{G_m{}^2}\right) \tag{8}$$

The operating transconductance, G_m, should be used in the formulas. In the case of jfets the G_m is specified at zero bias although the device is actually operated at a negative bias where the G_m is considerably less.

The gain of the rf amplifier should be just enough to overcome the losses of the tuned circuits and to override mixer noise. The rf passband may be anywhere from 60 to 150 kHz wide at 14 MHz,

even with multiple tuned circuits, and all signals within the passband will be fully amplified. An rf amplifier with a gain of 10 dB will reduce the effective dynamic range of the mixer by 10 dB for signals within the rf passband.

Select devices with high pinch-off handle less signal than the mixer, thereby degrading the performance of the receiver. Gain is preferably controlled by a manual or automatic attenuator between the antenna and the amplifying device.

Stability of the rf amplifier is a func-

table 1. Characteristics of selected rf amplifier fets. A 6BA6 vacuum tube is included for comparison purposes.

type	min G_m	device	max R_{eq}	mA I_{DSS}	max C_{gs}	max C_{rss}	min BV	V_P	ref freq	max NF
CP650	100,000	jfet	10	0.3—1.2A	25	25	25	2—10	50	2.5
40673	12,000	mosfet	55	5—35	6	.03	20*	—4	200	6
3N140	6,000	mosfet	94	5—30	7	.03	20*	—4	200	4.5
2N5397	6,000	jfet	167	10—30	5	1.2	25	1—6	450	3.5
2N4416A	4,500	jfet	222	5—15	4	0.8	35	2.5—6	100	2
E300	4,500	jfet	222	6—30	5.5	1.7	25	1—6	100	2
HEP802	2,000	jfet	500	2—20	-	-	25	-	-	-
6BA6	4,400	pentode	3550	-	5.5	.0035	-	-	-	-

* = BV_{DGO}, others BV_{GSS}

voltage, V_p for best linearity. Even within a given type it pays to select. The pinch-off voltage of the standard 2N4416A, for example, is specified at 2.5 to 6 volts. For best linearity jfets should be biased at approximately $V_p/2$. Dual-gate mosfets are run closer to zero bias, occasionally with slight positive bias.

Maximum linearity also demands that the operating point of the device be fixed and not varied for gain control purposes. The cross-modulation versus gain reduction characteristics of the 3N140 mosfet are shown in fig. 7.[11] In this case gain is reduced by varying the bias on gate 2. When operated at its normal conditions the device will handle a 130-mV signal before cross-modulation occurs. Changing gate-2 bias to reduce the gain by 5 dB results in dropping the signal-handling ability to 75 mV. Reducing the gain still further, to -15 dB, increases the signal-handling ability to 300 mV.

All devices, tubes, bipolars and fets, have linearity-vs-gain reduction curves similar to the 3N140. The exact location of the peaks and valleys varies from type to type and even within the same type. If you apply manual or automatic gain control to an rf amplifier it is possible to bias the device to a point where it will

tion of the feedback or reverse transfer characteristics of the device. This is composed almost entirely of capacitance, C_{rss} or C_{gd} in an fet. In the triode jfet C_{rss} runs from 1 to 3 pF, too high to use the device in an unneutralized common-source circuit. The dual-gate mosfet may be used in common-source circuits because its C_{rss} is only 0.03 pF. This is comparable to that of many pentodes although it is still considerably greater than the 0.0035 pF of the old reliable 6BA6.

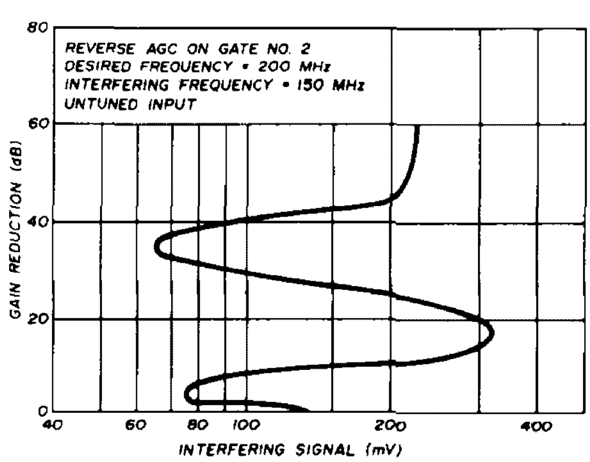


fig. 7. Level of interfering signal required to produce cross-modulation in a 3N140 dual-gate mosfet as gain is varied by changing the bias on gate 2.

Pentodes were long the workhorse of rf amplifier circuits so it is fitting to look at the pentode circuit before going on to newer devices. The 6AK5 came out of World War II research into better tubes for vhf. In 1946 it was used in the famous R9'er, a broadband, impedance-matching rf preamplifier (fig. 8).

The biasing and supply circuitry of the R9'er, a broadband, impedance-matching rf preamplifier (fig. 8).
rather than by a cathode potentiometer. The input and output tuned circuits are broad-banded by loading them down to a Q of 50 with resistors. Capacitive dividers permit adjustment of input and output impedance matches. The R9'er provided a dramatic improvement in noise factor

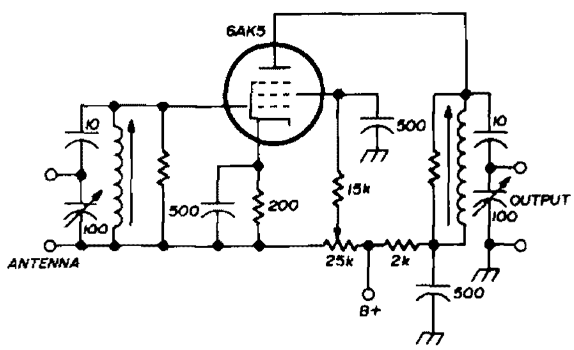


fig. 8. A pentode is used in the famous R9'er impedance-matching preamplifier popular in the late 1940s.

when connected ahead of some of the prewar receivers which typically had a NF of 14 dB or worse. The NF of the R9'er was 6 dB.

The modern day rf amplifier workhorses are the dual-gate mosfets such as the 3N140 and the 40673. The NF of the mosfet in the circuit of **fig. 9** with input matching optimized for noise, is 2 dB at 28 MHz and the gain is 26 dB.[12] Since both of these figures are better than required for high-frequency work, the excess gain and NF could well be traded for better signal-handling characteristics.

In this circuit resistors R3 and R4 form a voltage divider to provide 4 volts bias to gate 2 of the mosfet. Unlike tubes

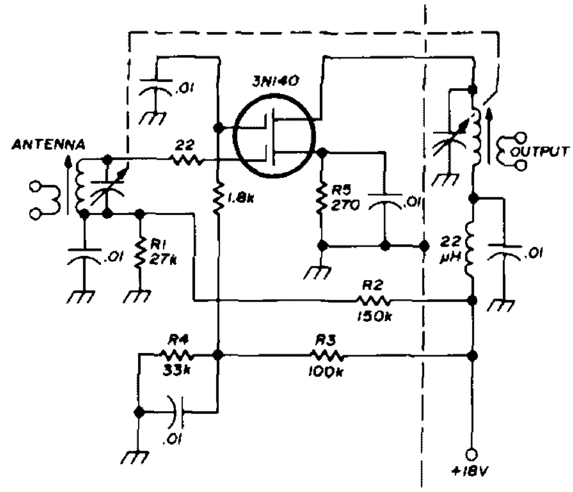


fig. 9. A modern rf amplifier stage using a dual-gate mosfet. Gain of this stage is 26 dB at 28 MHz; NF is 2 dB.

and jfets, the mosfet can be run at zero or positive bias and the transconductance is maximum near zero bias. Resistors R1 and R2 form another voltage divider to develop sufficient positive bias to overcome the negative bias of R5 and bias gate 1 near the zero bias point.

The jfet has such large feedback capacitance that, without neutralization, it is unstable in the common-source circuit. One remedy is to run the device in grounded gate. **Fig. 10** shows the circuit of a broadband preamplifier covering 0.5 to 50 MHz using a grounded-gate jfet.[13,14] The amplifier has a NF of less than 2.5 dB and a dynamic range – the range between the weakest usable signal and the signal which causes a 1-dB compression of the output–of 140 dB! The
pression of the output – of 140 dB! The
performance of this preamplifier is due

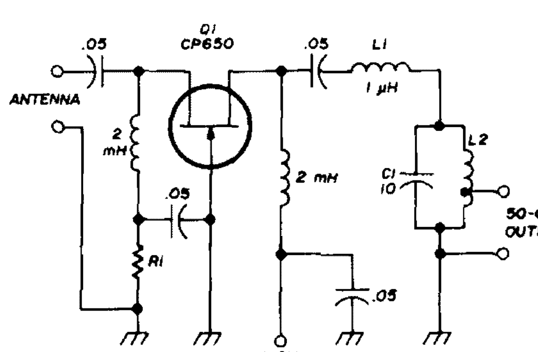


fig. 10. Wideband (500 kHz to 40 MHz) preamplifier using a power fet provides wide dynamic range.

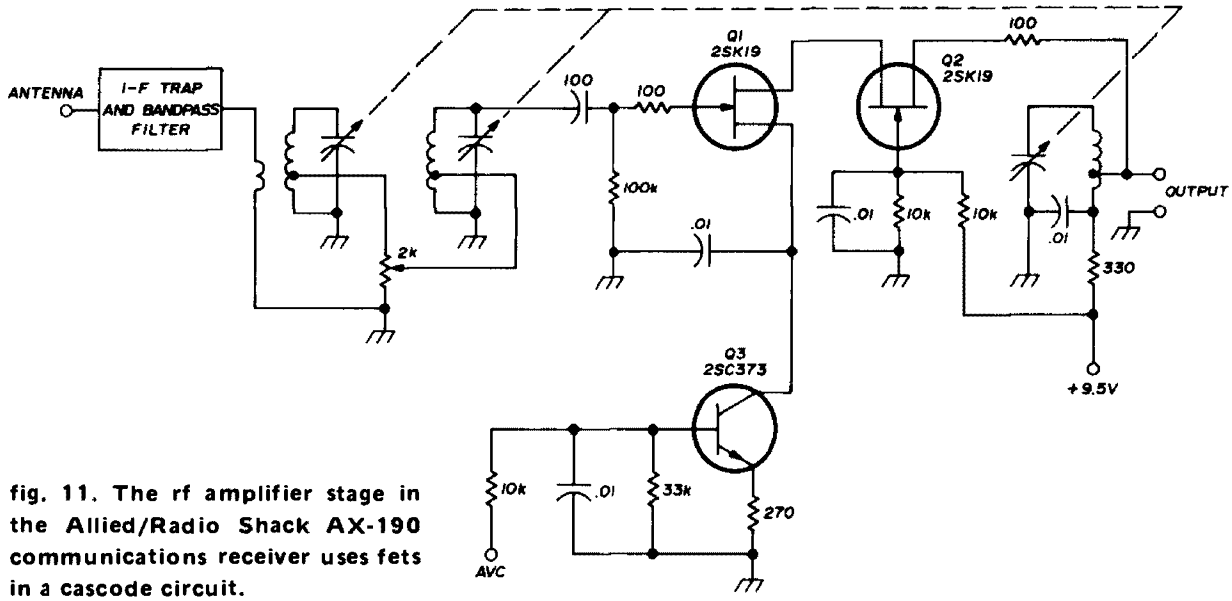


fig. 11. The rf amplifier stage in the Allied/Radio Shack AX-190 communications receiver uses fets in a cascode circuit.

power fet of the Crystalonics CP650 family. It consists of 15 fets in parallel on a single chip and provides a minimum G_m of 100,000 μmhos.

In fig. 10 resistor R1 is selected to set the drain current of Q1 at 30 to 40 mA at which point the G_m will be about 55,000 μmhos. The input impedance of a grounded-gate device is equal to approximately $1/G_m$ (18 ohms for this circuit). The input circuit results in a mismatch to a 50-ohm line but results in only about 1-dB loss in gain and improves the NF over the matched condition by approximately 0.5 dB.

The output circuitry is a lowpass filter, L1/C1, which provides a constant 200-ohm load to the fet over the entire frequency range. L2 is a 200- to 50-ohm matching transformer.

Jfets can also be used in the cascode circuit without neutralization. This circuit was chosen by the designers of the Allied/Radio Shack AX-190 receiver (fig. 11). The cascode circuit is stable because the input cannot "see" the output through the very small drain-to-source capacitance of the second device which uses the grounded-gate configuration.

Two tuned circuits are used in front of the amplifier, coupled through an attenuator pot. The drain of Q2 is tapped down on the output tank to preserve its Q. The 100-ohm resistors are parasitic suppressors. Transistor Q3 acts as a variable source resistor which is actuated by

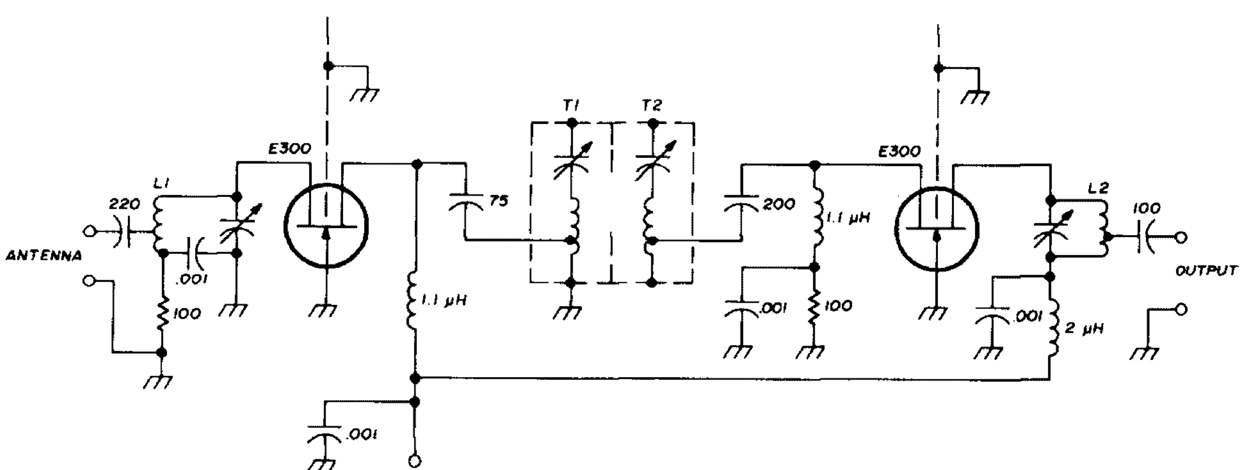


fig. 12. A selective rf amplifier using two grounded-gate fet stages. Selectivity is provided by the helical resonators T1 and T2. For high-frequency operation it would be better to tap the fets down on L1 and L2 to obtain more selectivity.

the avc system. This is one method of overcoming the incompatible avc requirements of fets and bipolar transistors.

Another interesting rf amplifier circuit which uses grounded-gate fets as buffer and impedance-matching devices for a pair of helical resonators is shown in fig. 12.[15] Designed for color tv reception on channel 10 while rejecting channels 9 and 11, it is claimed to provide 15-dB gain and a 4-dB NF at 193 MHz. The input and output circuits provide little selectivity because of the heavy loading of the fets. At high frequencies it would be better to tap the fets down on the tuned circuits so they could contribute to the overall selectivity of the receiver.

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ham radio

design data for pipe masts

One of the best materials for antenna masts is steel pipe — here is how to design your own

R.P. Haviland, W3MR, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

One of the best materials available for building self-supporting antenna masts is steel pipe. It is widely available, uniform in quality and reasonable in price. A well designed mast is adequately strong, neat and attractive, and relatively light weight. And, using steel pipe, it is not too difficult to design a fold-over mast which allows all antenna work to be done at ground level. Even maintenance on the mast itself does not require work at any great height.

However, attaining all of these advantages does require some design work. This is particularly important for safety. The purpose of this article is to present a set of design curves which will give a safe and satisfactory design while using the minimum of material.

construction

The general construction of a typical fold-over pipe mast is shown in **fig. 1**. At the top is the antenna and rotator, carried by the smallest size pipe. This is inserted into the upper end of the next size pipe for a short distance, and fastened by through-bolts or welding. The second section is inserted into the next larger, and so on. The bottom section is hinged to a fixed upright pipe, which gives the fold-over feature. It, in turn, nests into a larger section of pipe set into the ground. A yoke is provided to fasten the mast to the upright after erection. **Fig. 1** shows a block and tackle for pulling the mast to the vertical position but a winch, fastened to the upright, may be used instead.

Most mast designs use the widely available standard-weight pipe, each size of which nests neatly into the next larger size, over the range from 1½ to 4 inches (38 to 100 mm). Larger sizes still nest,

Standard and *Extra Strong* (ASTM nomenclature) are the two pipe weights commonly encountered. The American Petroleum Institute has a separate designation for well casing, but this is called tubing rather than pipe, although some sizes are identical to pipe sizes. The critical dimension for *Standard* weight pipe are:

size	OD	wall thickness
4-inch (102-mm)	4.5" (114 mm)	0.237" (6.0 mm)
3½-inch (89-mm)	4.0" (102 mm)	0.226" (5.5 mm)
3-inch (76-mm)	3.5" (89 mm)	0.216" (5.5 mm)
2½-inch (64-mm)	2.875" (73 mm)	0.203" (5.0 mm)
2-inch (51-mm)	2.375" (60 mm)	0.154" (4.0 mm)

The ASTM recommended fiber stress values for *Standard* weight pipe is 20,000 psi (bending). The design procedure presented here uses a 10% reduction from this stress figure, based on good used pipe. **Editor**

but there is a gap between the walls. Very high masts, or those with unusually heavy top loads, can be built with extra-strong or double-extra-strong pipe, but such designs are not considered here as the data are calculated for standard-weight pipe.

design criteria

The critical or design load on a section may be caused by wind load when the mast is vertical, or by erection load as the mast is being raised. Both loads should be calculated and the design chosen for the worst of the two.

For wind load, two design winds are commonly used. For most of the country, it is assumed that the worst wind to be encountered is 85 mph (137 Kph), a value to be expected once in 50 years or so. For Florida, the Gulf Coast and locations such as Cape Hatteras, a maximum wind of 125 mph (201 Kph) is also used. Your county engineer can provide the recommended value for your location (see reference 1).

During erection there is some deflection, or bending, of the mast. The greatest load occurs as each section is horizontal and this is the loading which must be designed for.

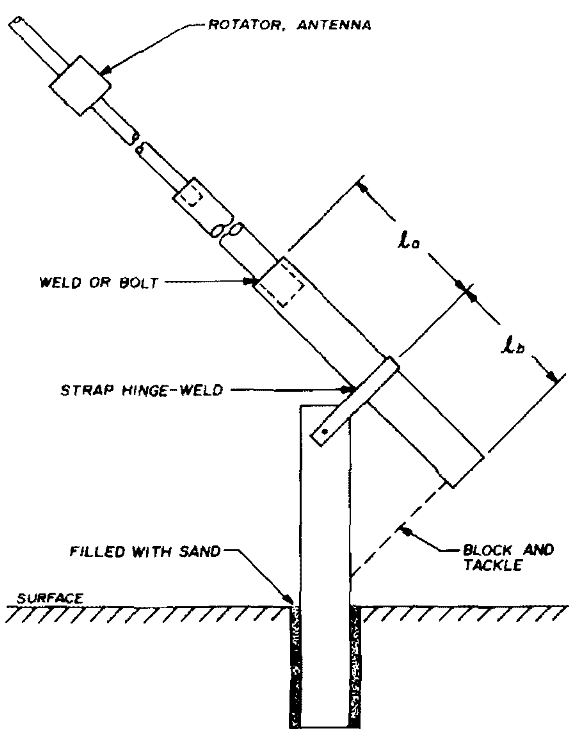

fig. 1. General layout of the fold-over pipe mast (not to scale).

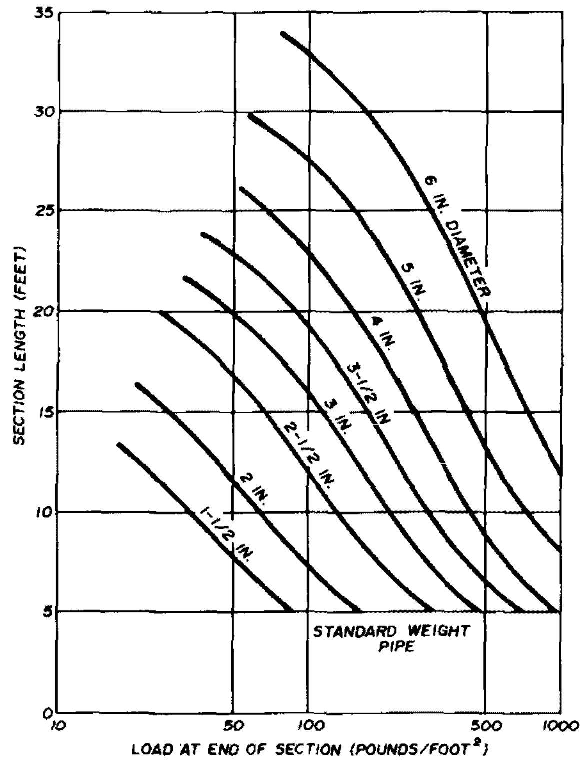

fig. 2. Allowable section length at erection for standard weight pipe, fiber stress = 18 kips.*

The wind and erection impose two different types of load on the section. One is the concentrated load at the topmost end of a section due to the forces on the section above. The second is the distributed load acting along the length of the section. As the concentrated load becomes larger there is less strength left for the distributed load, so the section length must become smaller. Accordingly, the problem of design is to determine the allowable section length.

The concentrated load during erection is the weight of the antenna, rotator and sections above the one being considered. The concentrated wind load includes the sum of *all* wind loads above the section being considered. The usual load is calculated on the basis of projected area. This is the area covered by the shadow of the

*The units of force are pounds, tons, kilograms, etc. In engineering practice the word *kip* is frequently used; it merely means a thousand pounds. Thus 18 kips can also be written 18,000 pounds. **Editor**

object. If the object is not symmetrical, as a Yagi beam, the largest projected area is used. The loading depends on whether the object is flat or round, as follows:

	wind loading pounds per square foot (kg per square meter) 85 mph (173 kph) wind	125 mph (201 kph) wind
Flat objects	30.3 (147.9)	65.9 (321.8)
Round objects	18.1 (88.4)	39.0 (190.4)

The projected area is often given in the instructions for commercially made antennas and rotators. It is easily calculated from the dimensions of the element.[1]

Given this concentrated load on the topmost section, design of the mast proper involves solving section load equations for allowable section length. To simplify this process the equations have been reduced to a series of graphs, **fig. 2** and 3 for load during erection, and figs. 4 and 5 for wind loads. Use of these curves will be explained through an example.

example

Assume that the design is for an all tubing six-meter antenna, having two square feet (0.186 square meter) projected area and weighing 15 pounds (6.8 kg). A small TV rotator is available, having one-half square foot (0.046 square meter) of mostly flat plate area, and weighing 8 pounds (3.6 kg). This area is not subjected to unusual winds. Mast height is forty feet (12.2 meters).

The concentrated load on the top section is 15 plus 8 or 23 pounds (10.4 kg). Entering fig. **2** at the bottom with this weight and moving upwards, it is seen that the top section could consist of 12-feet (3.7-meters) of 1½-inch (38-mm) pipe, or 16-feet (4.9-meters) of 2-inch (51-mm) pipe or 20-feet (6.1-meters) of 2½-inch (64-mm) pipe. In keeping with the scale of the antenna, suppose the 1½-inch diameter (38-mm) pipe is used.

The concentrated wind loading is due to 2 square feet (0.186 square meter) of antenna and one-half square foot (0.046 square meter) of rotator. From the table above, the loading is 2 x 18.1 plus 0.5 x 30.3, or 51 pounds per square foot (249 kg per square meter). Reading upward from this load on fig. 4, it is seen that the maximum allowable length for 1½-inch (38-mm) pipe is 8 feet (2.4 meters). Since this is the critical value, it becomes the length of the topmost section.

Assume that the sections are to be fastened by welding, with 6-inch (15.2-cm) insertion into the next section. From fig. 3, the weight of the 8½-foot (2.6-meter) total of the top section is 23 pounds (10.4 kg). The wind loading on the exposed 8 feet (2.4 meters), from **fig.** 5, is 25 pounds per square foot (122.1 kg per square meter). Thus, the weight load at the top of the second section is 23 + 23 or 46 pounds (20.9 kg) and the wind loading is 51 + 25 or 76 pounds per square foot (371.1 kg per square meter).

Using fig. **2** again, the maximum allowable length of the next section using the nesting 2-inch (51-mm) pipe is 11½ feet (3.5 meters) for erection loads. From fig. 4, the allowable length for wind loads is 9 feet, which becomes the section length. Proceeding as before, the loads on the next section are 46 plus 35 or 81 pounds (36.7 kg) during erection, and 76 plus 35 or 111 pounds per square foot (541.9 kg per square meter) for wind.

Again, using fig. **2** and 4, the allowable length of 2½-inch (64-mm) pipe is 13 feet

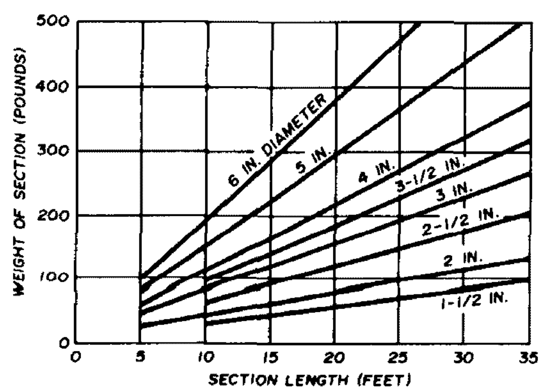

fig. 3. Weight of standard pipe.

(4 meters) for erection load, and 12½ feet (3.8 meters) for wind load. The 12½ feet (3.8 meters) is the length ℓ_a in **fig. 1**. The load on the section ℓ_b in **fig. 1** is the same in magnitude, so this part could also be 12½-feet (3.8-meters) long. However, a stock length for pipe is 21 feet (6.4 meters). Assume that this is all that is available. Then the third section will need to end one-foot (30-cm) above ground to reach the desired 40-feet (12.2-meters) total height. This is not unreasonable.

10½ feet (3.2 meters) of the lower section plus some amount on the upright. Assume that the upright is fully exposed, a safe assumption. The wind load to the top of the upright is 111 plus 55 or 166 pounds per square foot (810.5 kg per square meter) maximum, the exact value depending on the final choice of upright length. From **fig. 4**, the upright can be only 6-feet (1.8-meter) long if it is 2½-inch (64-mm) diameter, or 13-feet (4-meters) long if it is 3-inch (76-mm)

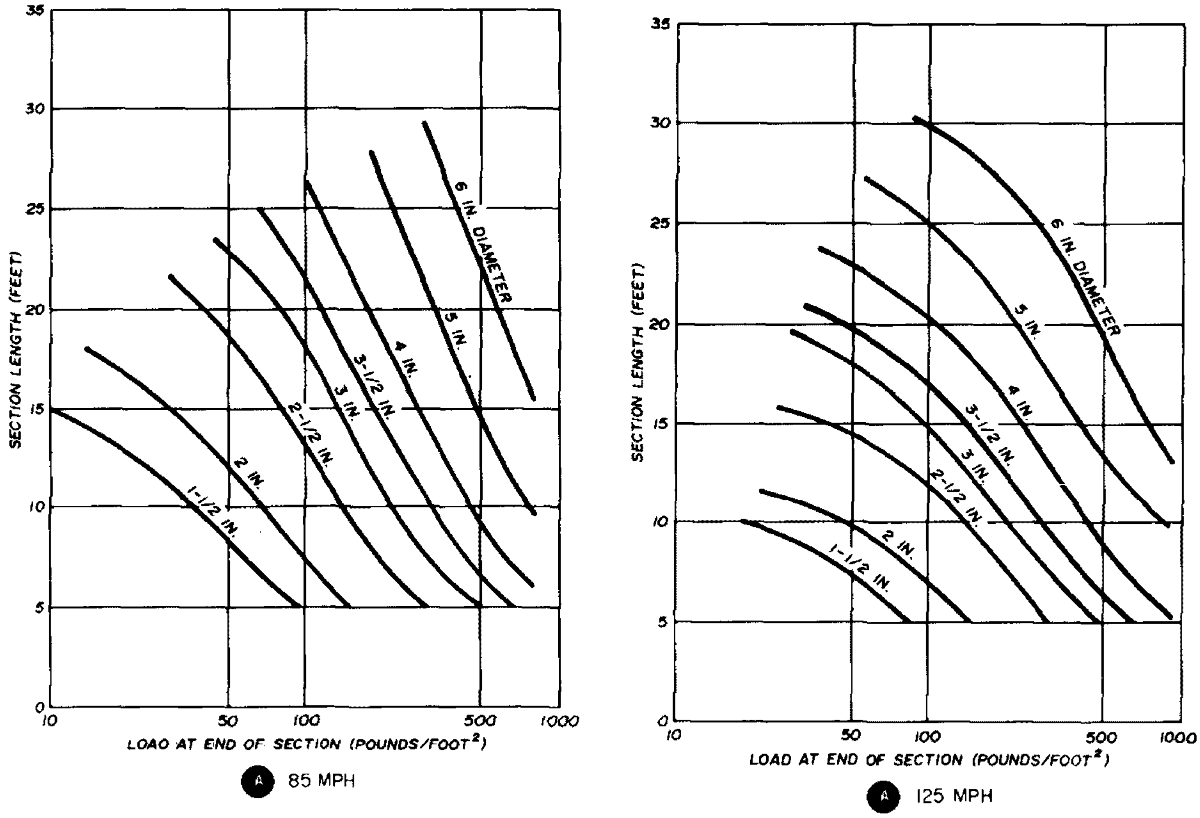

fig. 4. Maximum allowable section length for standard weight pipe with winds of 85 and 125 mph (fiber stress = 18 kips).

If a counterweight is added to the lower part of the third section to just balance the top weight, the erection loads on the fixed upright pipe are essentially zero. Even if no counterweight is used, the balancing effect of the part ℓ_b of **fig. 1** reduces the load on the upright to less than the load on section ℓ_a of **fig. 1**. Thus, *if the upright is no smaller than the lowest mast section, it will have adequate strength for erection.*

The wind load on the upright is that of the upper sections plus that on the top

diameter. Since 12½ feet (3.8 meters) is needed as a minimum, this is just about right (half of the 21-foot [6.4-meter] length of the 2½-inch [64 mm] section, plus one-foot [30-cm] ground clearance).

Factors affecting the length of pipe buried in the ground are discussed below. For this example, assume that this is ten percent of *mast* height, or 4 feet (1.2 meter). Total upright length is thus 13½ plus 4 or 17½ feet (5.3 meters). The *jacket section buried in* the ground needs to have one-inch (25-mm) clearance, so

it needs to be a four-foot (1.2-meter) length of 5-inch (127-mm) diameter pipe.

The results of this design example are:

Top section: 1½-inch (38-mm) diameter top section, total length 8½ feet (2.6 meters), exposed 8 feet (2.4 meters).

Second section: 2-inch (51-mm) diameter second section, total length 9½ feet (2.9 meters), exposed 9 feet (2.7 meters).

Lower section: 2½-inch (64-mm) diameter lower section, total length 21 feet (6.4 meters), hinge at 12½ feet (3.8 meters), 1-foot (30-cm) ground clearance at bottom.

Upright: 3-inch (76-mm) diameter upright, total length 17½ feet (5.3 meters), exposed 13½ feet (4.1 meters), buried 4 feet (1.2 meters).

Jacket: 5-inch (127-mm) diameter jacket, total length 4 feet (1.2 meters), all buried.

If necessary, this design could be carried higher, using larger pipe sizes.

It is often necessary to try several initial assumptions as to length and diameter of the top section. With a little practice, this can be done in a few minutes.

foundations

Because of the great variability of soils, it is not possible to provide a set of all-purpose design curves for these. The best way of proceeding is to work with your county engineer, and use the practices developed for your particular area. The local power or telephone company should also be able to supply the necessary data.

For reasonably good soils, such as firm loams or clays, a good starting point is to assume that the foundation depth is equal to ten percent of the height, with the jacket set in concrete of sufficient size to keep the soil load to a safe value. A maximum load of 4000 pounds per square foot (19530 kg per square meter) is often used, with the design being adjusted to give 100% safety factor above the design load. If you haven't done this work before, the county engineer can show you the steps.

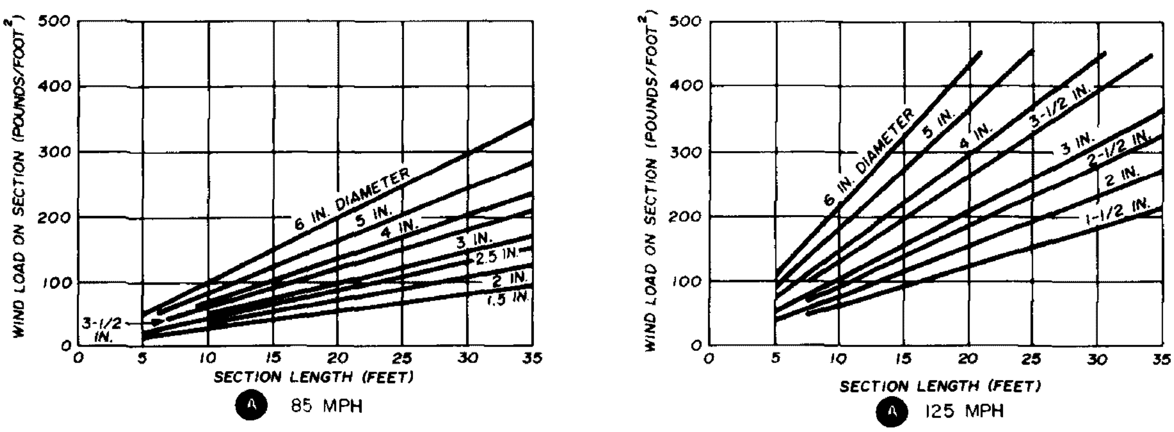

fig. 5. Wind loading for standard weight pipe, 85- and 125-mph winds.

safety

Any antenna mast can become a hazard if good safety practices are not followed. Remember that a quarter- or half-ton of steel thirty- to seventy-feet (9- to 21- meters) in the air is no toy. If you lack experience or don't have the proper facilities, get qualified help. Always remember, *safety is no accident*.

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reciprocating detector converter

By popular demand — a refinement of previous reciprocating-detector circuits appearing in *ham radio*

Stirling Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

In my last article on the reciprocating detector[1] the final paragraph contained the statement, " . . . It is hoped that other amateurs will try it and perhaps find some of the features we missed; or perhaps shoot down those reported." In the months following publication of the article, I have received considerable mail explaining what I missed, or telling me I was copying some one else's idea, which had been invented using tube circuits in 1907. For the most part, the questions or descriptions were useful in answering all who posed other questions. But most of all, demonstration of several shortcomings were made obvious. Acknowledgements usually follow the last paragraph in an article; but mine must come first, and I thank all who wrote. This article is presented to answer most of the queries received and to set the record straight on what the reciprocating detector (RD) is, and what it is not.

typical queries

Many requests were for circuitry that could be used to incorporate the RD in receivers with i-f amplifiers at frequencies well above the limits of the practical reciprocating detector. The method of obtaining the RD reference is through feedback, which is partly carrier-level controlled and which would require

shielding of the adjacent circuits and the use of semiconductors well out of the price range of the average amateur. Other requests desiring to use the RD with transceivers required extensive investigation into the operation of these sets. In most cases I recommended that the amount of work would not justify the results.

older sets

Many amateurs installed the RD in older sets that had single filters, or none at all, with the hope that ssb, as well as the other features described would result in a receiver that could compete in the melee on bandedge pileups, or that could put these receivers in the same class with a modern receiver. This is not possible, and it was not my intention to mislead anyone into thinking it was.

The problem here is perhaps best resolved by careful consideration of the facts on how the receiver operates; for some because of the avc characteristics or the lack of them.

The set could never be used successfully on ssb. Why? Let's take the case of a receiver with single-filter selectivity. Suppose the receiver has one filter 3.1-kHz wide with a 455-kHz i-f. A reciprocating detector with a 455-kHz reference frequency will require the user to offset-tune the front end. He will receive some of the sideband he is looking for and will have all the pertinent features of noise impulse suppression of the RD, but part of the sideband will be clipped and will sound hard.

If there is no filter at all, and the bandwidth is very wide (which is usually the case), then either sideband will get through easily, along with the adjacent channel. Obviously filters are a *must* and dual filters are the best in my opinion.

The attack characteristics of the avc in older sets are such that these systems cannot be used with the reciprocating detector. The RD can't stand overloads, and in many cases the rf signal presented by the i-f amplifier of older sets to the RD is in excess of 3 volts, which will certainly saturate it. I don't have a cure for this problem in the design presented here. Unless you can redesign your old receiver to accommodate today's signal amplitudes, I can't recommend an RD to replace your existing detector unless you wish to use it for some of the other modes of operation.

Before describing the RD converter, I'll describe the operation of the RD and define some terms.

circuit description

Let's start by looking at the circuits of **fig. 1**. The description deals first with the circuit of **fig. 1A**, a bit different from most detector circuits. A dsb signal will be used to actuate the RD switch. The rf input is fed into a battery and resistor combination, which provides bias to keep transistor Q3 just below, but not quite at, cutoff. Transistor Q3 therefore serves as a half-wave rectifier for the incoming signal and as a current source for the emitters of Q1, Q2. This current source is the reference for the base of Q2, enabling it to conduct on the positive reference peaks. Therefore, since Q1 and Q2 emitters are driven from a common-current source, Q1 is enabled and conducts when Q2 is off, but only on negative reference peaks. Neither Q1 nor Q2 can conduct unless Q3 is also enabled, which occurs only on positive rf signal peaks. Therefore Q2 conducts when the rf signal is in phase with the reference signal, and Q1 conducts only on the 180-degree difference with respect to the reference.

Now, since the audio and rf outputs are obtained differentially from those two switching transistors, a transfer of the conduction of Q1 to Q2, or vice versa, causes a polarity flip. It is this flipping action that is required to convert the waveform to a sine wave.

In the paragraphs above I have

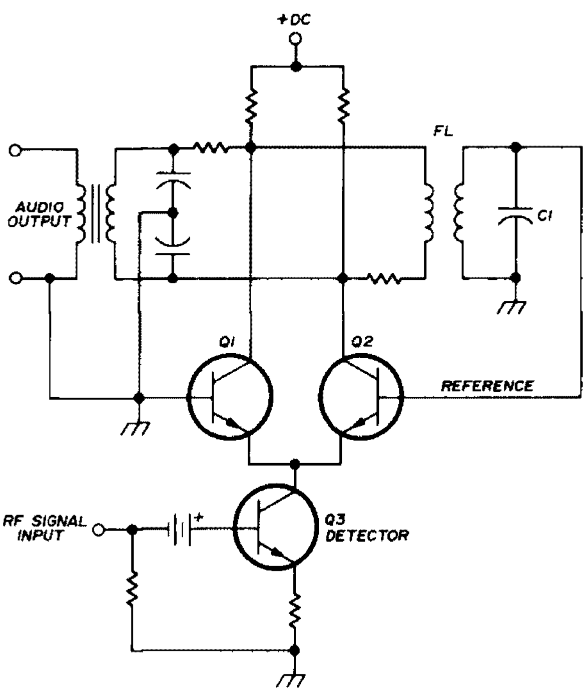
continually used the word "reference," and in doing so may have confused the issue. By reference, I mean the beat signal that is synthesized from the rf input signal. Through feedback, routing the signal through the filter, FL, we have automatically provided a beat frequency oscillator, which is dependent upon the signal. More about this later.

synchronous mode

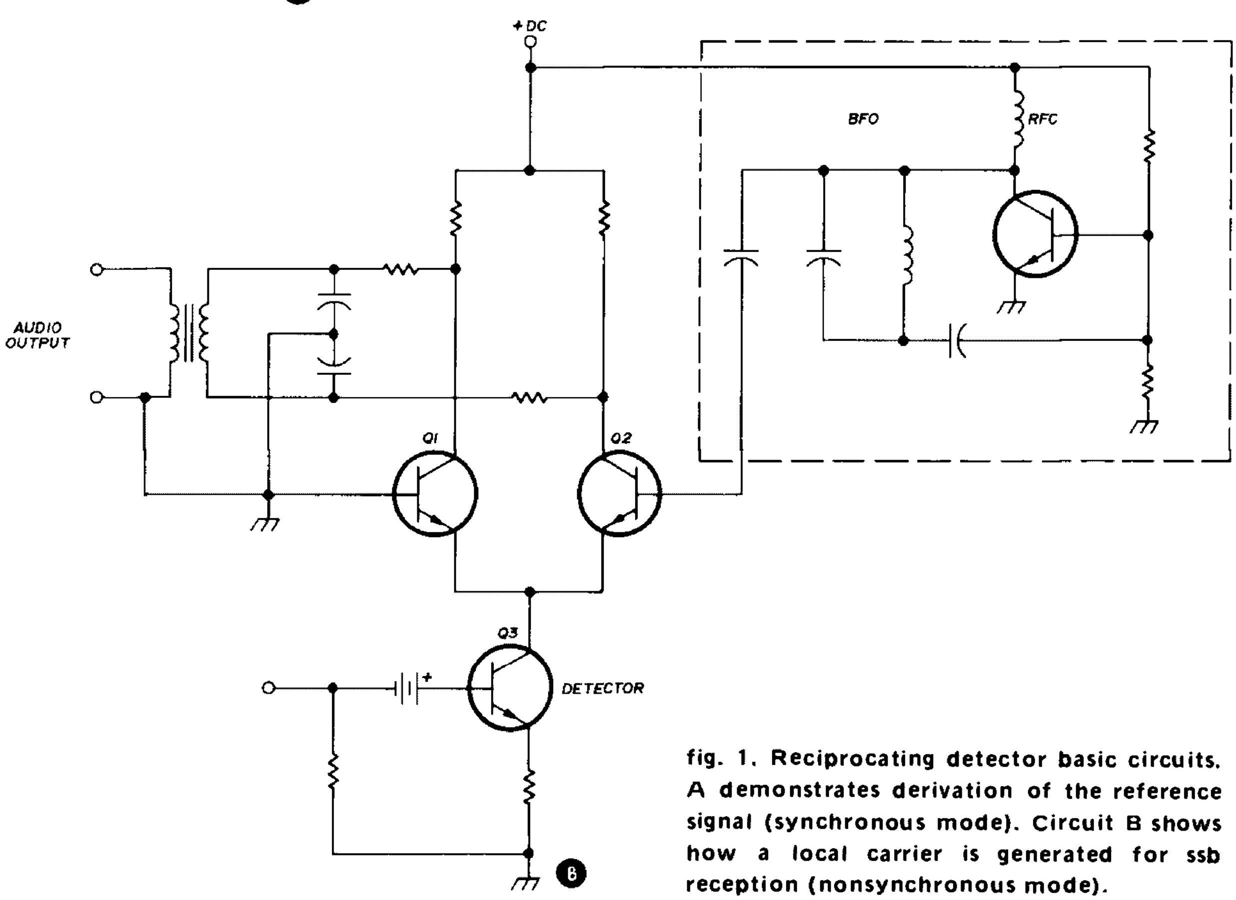
So far I've talked about a system that is self-controlled through the use of the carrier-synthesized reference. Now comes another term we must use: the synchronous mode. The reference signal is synchronous with the carrier because it is *generated* by the carrier. The feedback through the filter, FL, establishes this action. The signal is recovered from the switch through an audio transformer (acting as a low-pass filter in the demonstration circuit).

nonsynchronous mode

Since it has been demonstrated how the reference signal is derived from the received (carrier) signal, and you now know what is meant by the synchronous mode, I'll try to answer the question most amateurs get around to asking: "How do you get a reference when you have a single sideband signal with no



A



B

fig. 1. Reciprocating detector basic circuits. A demonstrates derivation of the reference signal (synchronous mode). Circuit B shows how a local carrier is generated for ssb reception (nonsynchronous mode).

carrier?" A third definition now rears its head: nonsynchronous mode—because there is *no* carrier-generated bfo signal if the carrier is properly suppressed.

It will be easier to understand what's going on with the aid of **fig. 1B**. It's the same circuit as in **fig. 1A** but without the feedback loop. It uses a separate oscillator *in* the same fashion as a product detector. It is, in fact, exactly like a product detector.

Let's rearrange some of the circuits and voltages and see what happens. First we replace the input signal to transistor Q3 with a small dc voltage. This voltage is converted to a fixed current in the collector of Q3. Now we replace the bfo with the filter circuit. Transistors Q1 and Q2 are now part of a differential amplifier. Through regenerative feedback Q1 and Q2 form a simple oscillator operating at the filter center frequency.

The dc voltage impressed on the base of Q3 is the average value of the half-wave rectified signal. The flipping action previously described for the synchronous mode still occurs. However, it's difficult to follow. This is because of the intermediate phase changes, which result from the frequency difference between the input signal and the self-generated reference signal. In this mode, the reference level is no longer completely amplitude controlled by the input signal, but it does have signal-induced phase fluctuations. This signal is now pure rf since it is the filtered version of the signal envelope. So now the question should really have been, "How do you detect a reference for ssb signals?"

bandwidth and noise reduction

We will now discuss the lock-in range and filter bandwidth, which many readers want to decrease because of the noise-elimination effects. Let's go back to dsb. The synchronous bandwidth of the signal is about one-third of the filter passband, or about 150 Hz. This is the case where phase correlation, previously discussed, is used. With an ssb input signal, the reference point follows the signal. Now, if the sides of the filter are too steep, phase synchronism will not be consistent and the reference signal will hop around trying to follow the input signal.

The peak deviation of a virtually unfiltered RD reference signal can't swing more than 30 degrees. This is not too obvious. Remember that the 180-degree phase change in the incoming signal is completely wiped out by the flipping action of the reciprocating switch. So a filter with a bandwidth of 500 Hz or so will yield little or no jitter, even if the input is random noise. As for amplitude variations, this same filtering reduces these changes in the reference signal to a degree that most amplitude variations that approach zero are not evident. The fluctuation rate of these variations could be in the vicinity of 250 Hz. It should now be obvious why the circuit can and does reject impulse noise of short duration and reduces the intensity of long static crashes. It should also be obvious why a filter with sides too steep or too narrow will not help, except, perhaps, on noise. If, however, the filter is too narrow and only improves noise rejection, the phase *jitter will ruin* the effectiveness of the reciprocating detector for its other purposes.

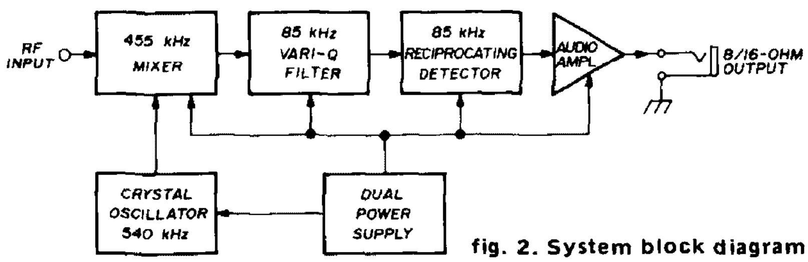


fig. 2. System block diagram.

455-kHz i-f receivers

It took a lot of looking to find a way

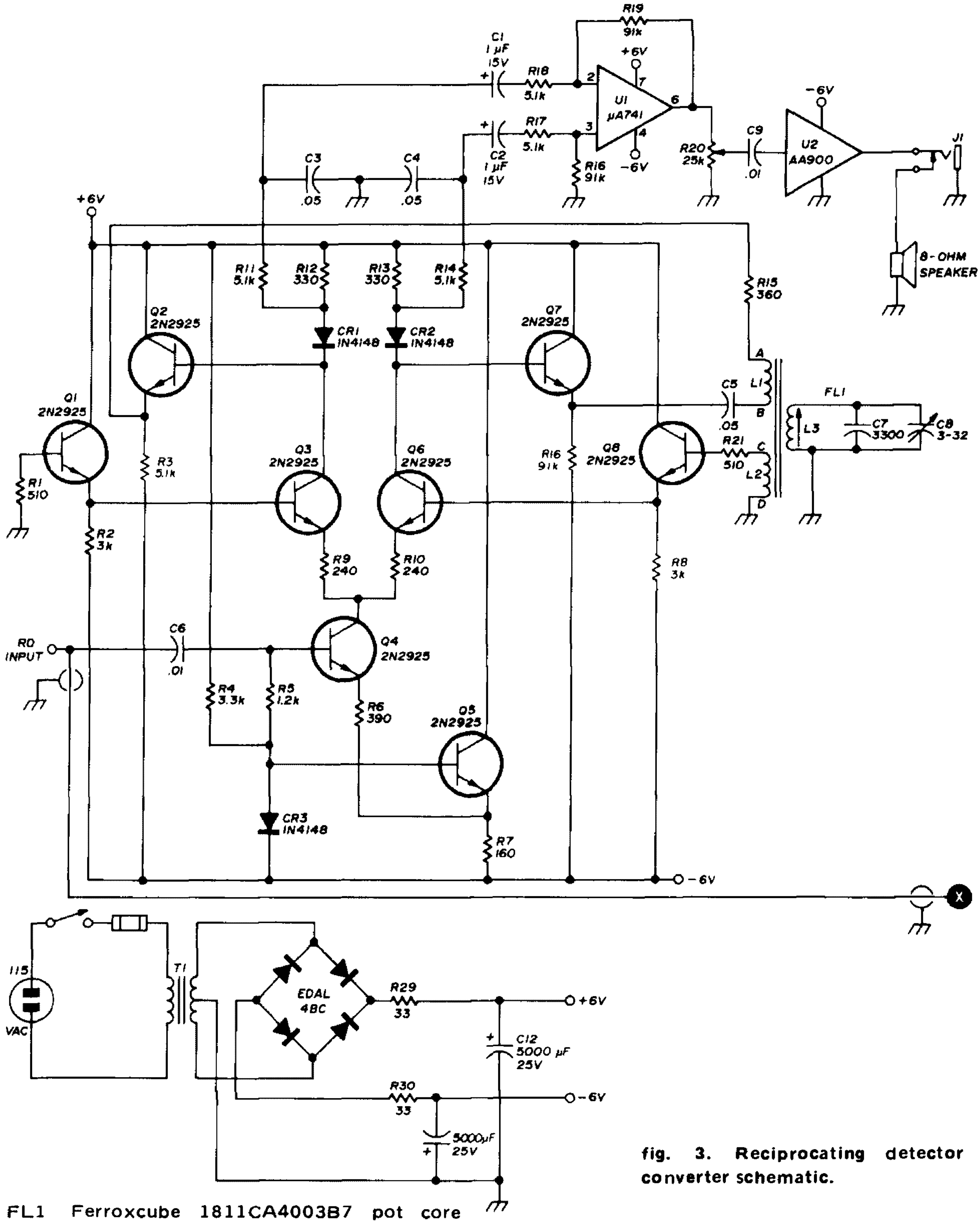
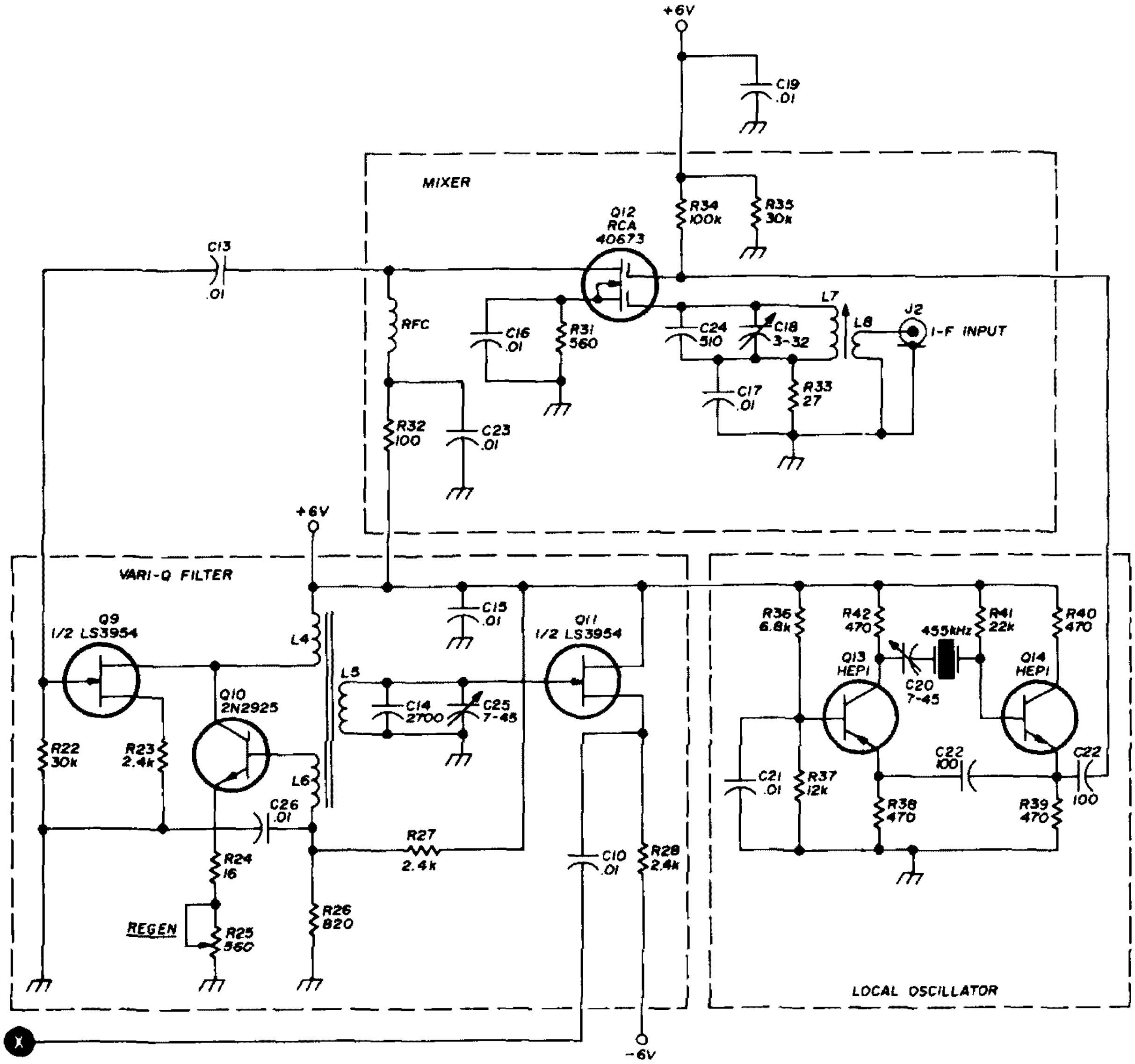


fig. 3. Reciprocating detector converter schematic.

FL1	Ferroxcube 1811CA4003B7 pot core with 1811 bobbin. L1 is 5 turns no. 30 enamelled; L2 is 36 turns no. 30 enamelled; L3 is 30 turns no. 36 enamelled
L4, L5, L6	Vari-Q filter wound on Ferroxcube 1811-CA250CB7 pot core with 1811 bobbin. L4 is 15 turns no. 30 enamelled; L5 is 130 turns no. 36 enamelled; L6 is 26 turns no. 30 enamelled
L7, L8	Mixer coil, 455 kHz. L7 is 130 turns no. 36 on Amidon Associates T44-15 toroid core. L8 is 8 turns no. 36 wound on same toroid

that would allow the RD to be used with almost any receiver (see fig. 2). A mixer is modulated by a crystal-controlled oscillator. The mixer input has a tuned circuit adjusted to match a 455-kHz i-f output. It can easily be coupled to the last i-f stage through a coaxial cable and a small capacitor. The input circuit can be adjusted to accept a wide range of frequencies simply by changing the input transformer



to one for the desired frequency. The crystal must also be changed.

The vari-Q filter is excellent for reducing interference. Its passband may be set by a variable capacitor and adjusted by a potentiometer (C15 and R25, respectively, fig. 3). Adjustment is variable over a 15-kHz range.

construction

The circuit (fig. 3) is conventional. The reference filter includes a variable capacitor operated from the front panel. This control allows the reference signal to be offset for receiving ssb signals. The control can be marked to indicate upper or lower sideband.

*PC board available from the author. Write for details. Please include a self-addressed stamped envelope.

The circuit is constructed on a PC board* except for the af amplifier and power supply. My unit is contained in a 4x4x8-inch (10x10x20-cm) cabinet finished to match other equipment on the operating desk.

Point-to-point wiring is used. Components in the parts list and schematic are not sacred except for the pot cores. Semiconductors should be silicon. Almost any general-purpose npn transistors may be used. The mixer transistor is an RCA 40673, although substitutes are available that will work just as well.

references

1. Stirling Olberg, W1SNN, "Reciprocating Detector," *ham radio*, March, 1972, page 32.
2. Stirling Olberg, W1SNN, "Vari-Q Filter," *ham radio*, September, 1973, page 62.

ham radio

miniature filament transformers

When miniature filament transformers are used in low-voltage power supplies the output voltage may be higher than expected

John T. Bailey, 86 Great Hills Road, Short Hills, New Jersey 07078

In the old days when you bought a filament transformer you hardly ever bothered to measure the secondary voltage. You knew that, at no load, it would be roughly 10% or so higher than at full load. This allowed for a reduction when loaded and also allowed for some voltage drop in the connecting wires to the tube sockets. Also, filament transformers were used primarily in those days to heat filaments or heaters and these voltage requirements weren't very critical.

Now things have changed. With miniaturization and the lower power-supply voltages required for semiconductor circuits using transistors, ICs and LEDs in analog and digital equipment, filament transformers are often selected by the experimenter as convenient power-supply transformers. Sometimes this is disastrous because, unless the transformer is fully loaded to its rated current capacity, the secondary voltage may be much higher than expected.

power supply

Consider an experience I had. I needed a 15-volt dc power supply to furnish 20 mA to a transistor circuit. Good regulation was not a requirement but this transistor circuit absolutely could not tolerate a supply voltage greater than 20 volts. Anything over 20 volts could zap the works.

For the power supply I decided to use a 12-volt filament transformer and a full-wave bridge rectifier feeding a capacitive filter. My initial rough calculations indicated that the output couldn't possibly be more than 20 volts even at no load after allowing for a 110% no-load secondary voltage (plus an increase of 7% because my line voltage is a steady 125 volts ac rather than 117 volts). This worked out to 20-volts dc maximum at no load.

Since the load current of 20 mA was quite modest and since I wanted to hold the size of the power supply down, I looked around for a miniature 12-volt filament transformer with a lower current rating than the generally available 600-mA units. At Radio Shack I found a 12-volt, 300-mA transformer with a core that measured only 1 3/8 x 1 1/8 x 3/8 inches (35 x 29 x 10 mm). I hooked it up, connected my voltage-sensitive load, turned on the juice and promptly zapped the transistor. What went wrong? Transients? Not this time. I measured the no-load dc output of my power supply (belatedly) and found it to be 27.5 volts!

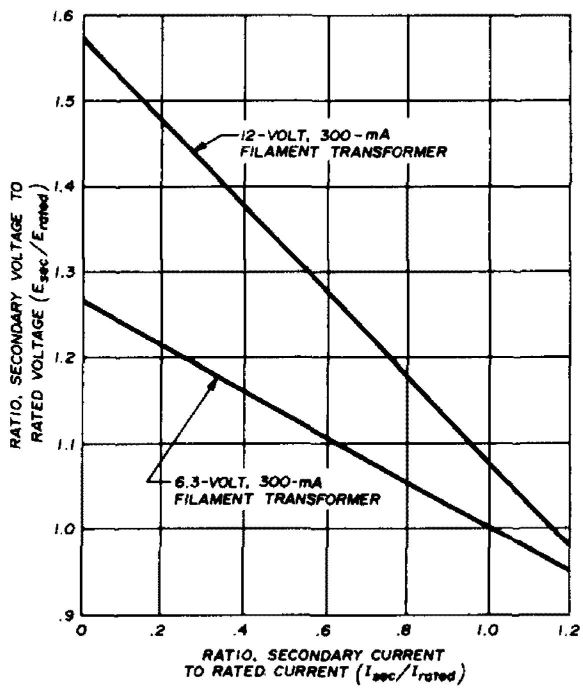


fig. 1. Chart shows how the secondary voltage increases as the secondary current is decreased in two miniature filament transformers.

The no-load secondary voltage to the bridge rectifier measured 20 volts ac instead of 14 volts as originally estimated.

miniature transformers

This sad experience aroused my curiosity about *miniature* filament transformers so I made some measurements and tests. I pass the results along for the benefit of others. I went back to Radio Shack and picked up one of their 6.3-volt, 300-mA filament transformers. This one had the same core size as the 12-volt, 300-mA unit even though the VA ratings have a 2 to 1 ratio. Now I had a 6.3-volt and a 12-volt transformer to test and compare. Each transformer was connected in turn to a Variac to energize the primary at the rated 117-volts ac. An adjustable resistive load was connected to the secondary and varied to draw from rated full-load to no-load secondary current.

The results are shown in fig. 1. From this graph it is evident why my equipment was zapped. The no-load or light-load secondary voltage is considerably higher than the voltage at full-rated current. This revelation isn't meant to be a condemnation of the Radio Shack transformers. Actually they are well-made, American-produced products which put out their specified voltages at their full-rated currents.

Sometimes when electronic components are miniaturized there are trade-offs involved. With transformers the core losses go up and the copper losses are greater due to the higher resistance of the small wire used. Would you believe that the primary current at no-load can be greater than when the secondary is fully loaded? I measured these primary currents on the 6.3-volt, 300-mA transformer. Using a thermocouple-type milliammeter to get an rms measurement, I read 49 mA at no-load and 42 mA at full load. Using a scope revealed a more distorted primary current waveform at no-load which confirms that core losses are high with miniature transformers.

Next I measured the resistances of the windings. The 12-volt unit had a primary resistance of 270 ohms and a secondary resistance of 8 ohms. The 6.3-volt unit had 420- and 3-ohm windings. These readings are many times higher than the corresponding resistance for physically larger transformers of higher current ratings and partially account for the low regulation factor. In conclusion, when using miniature filament transformers you must expect higher-than-normal voltages when less-than-rated secondary currents are being drawn.

ham radio

versatile squelch-audio amplifier for fm receivers

A simple squelch system with sharply defined threshold that may be added to any fm receiver

Robert C. Harris, WB4WSU, Box 487, Leesburg, Virginia 22075

Next to attaining the impossible is building a squelch-audio circuit that really works. Even more unlikely is locating a squelch circuit which will function well in a variety of different fm receivers. The circuit described here achieves the next to impossible with readily available components.

circuit

The audio amplifier is a conventional, run-of-the-mill IC amplifier. Others may work just as well. Audio from the fm detector output is connected to the audio gain control through shielded audio line, properly filtered and amplified to drive a loudspeaker. A second, similar arrangement connects the detector output to the squelch sensitivity control – the signal is then filtered by a simple RC circuit, amplified by Q1 and rectified. Only noise above the audio portion of the signal spectrum is affected by this processing. Under no-signal conditions negative bias is maximum and Q2 is turned off at a threshold level determined by the sensitivity control setting. Transistor Q2 then begins a logic toggling action through U1, a 7400 IC. A low on pin 8 of U1 clamps off a portion of U2, thus quieting the speaker. A signal carrier reverses this process, passing audio uninhibited

through U2 to the speaker. The toggle effect caused by the SN7400 IC provides an extremely sharply defined threshold.

The only limiting factor with this squelch-audio amplifier is the no-signal, noise output voltage from the fm detector which should be at least 0.75 volt ac as read on a sensitive voltmeter. Eliminating no-signal noise is one of the requirements of a good squelch system; replacing the SN7400 with a conventional Darlington transistor pair makes much lower clamping currents possible.

results

This squelch system compares favorably to the best of commercial circuits, and is readily adaptable to a multitude of fm receivers, unlike many commercial circuits. The circuit has served flawlessly

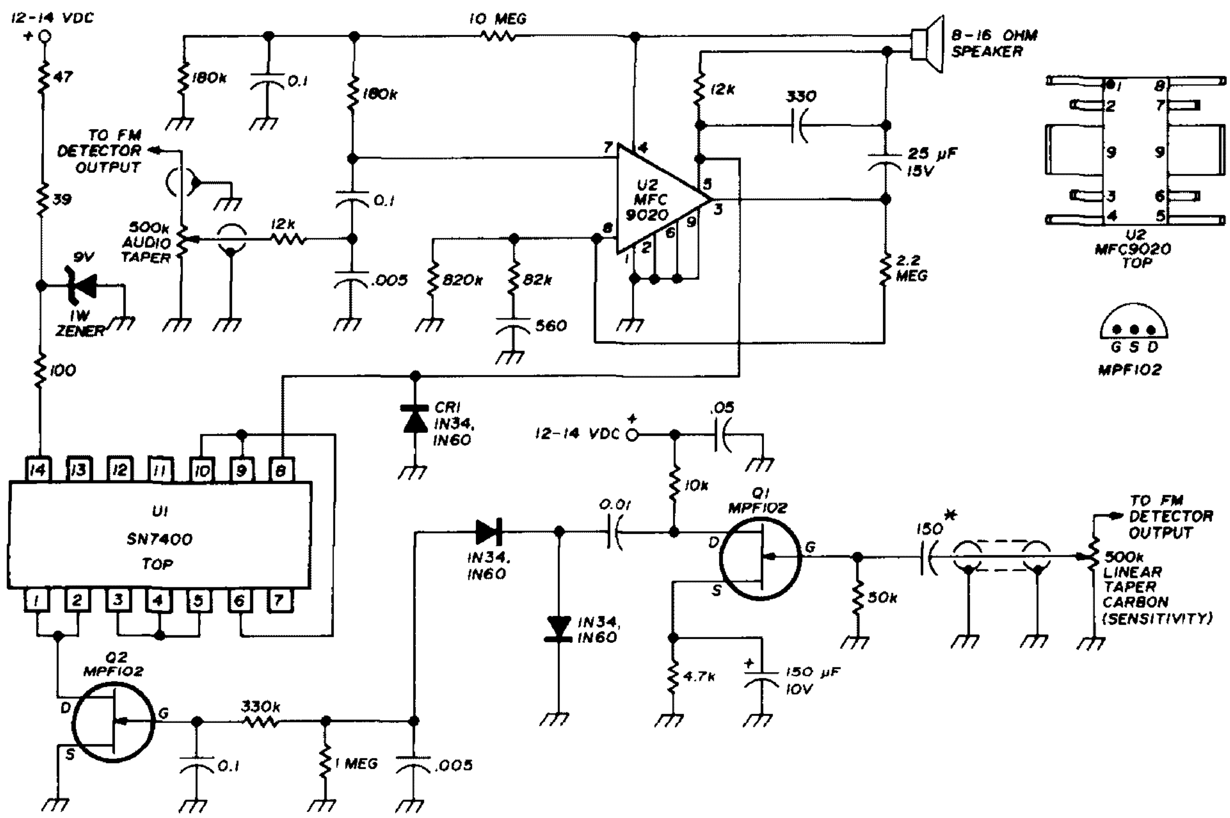


fig. 1. Schematic diagram for the versatile squelch-audio amplifier system. Diode CR1 may be eliminated if the B+ line is not keyed between transmit and receive. The 150-pF capacitor marked with an asterisk may be replaced with a smaller value, if necessary, for increased voice rejection.

the other is making it possible for very weak signals to pass. This circuit responds to the weakest of signals.

This squelch circuit performs well with supply voltages from 12- to about 14-volts dc. Squelch sensitivity is affected slightly by the varying power supply voltages inherent with mobile operation, but is well within acceptable limits. Current drain is on the order of 35 mA although lower values are possible through experimentation. For instance,

for two years as the replacement for an entire T-33 Motorola Dispatcher audio squelch system at a considerable reduction in current drain. Plenty of audio output is available and two hours of perforated-board construction is about all it takes to assembly the circuitry. The original was built on a 3x3-inch (76x76-mm) board and mounted inside the receiver. Nominal cost of the unit is about $5.00

ham radio

the ham notebook

the Heathkit HO-10 and SB-610 as an RTTY monitor scope

Any RTTY enthusiast who has tried to use either the HO-10 or SB-610 with the ST-series of terminal units has been quite disappointed. Both of these scopes offer too little gain for a good display with the one-megohm isolation resistors in the TU discriminator. These resistors prevent external cables (and scope) from loading the discriminator and also provide some filtering to clean up the display.

One solution that has been used is to build a pair of amplifiers into the TU. This is a good solution and is relatively simple. However, I have three TUs and only one scope devoted to RTTY monitoring. The obvious answer is to include the amplifiers in the scope rather than the TU. The scope, however, does not have the voltages necessary for a solid-state amplifier nor the room for an additional tube. There is a tube in the unit (both HO-10 and SB-610) which is never used at this station and is probably seldom used elsewhere. The two-tone generator uses a 6J11, dual pentode, to generate the tones. It is possible to use this oscillator stage as an amplifier with a minimum of additional parts and expense. The display is nearly as good as that provided by my lab scope.

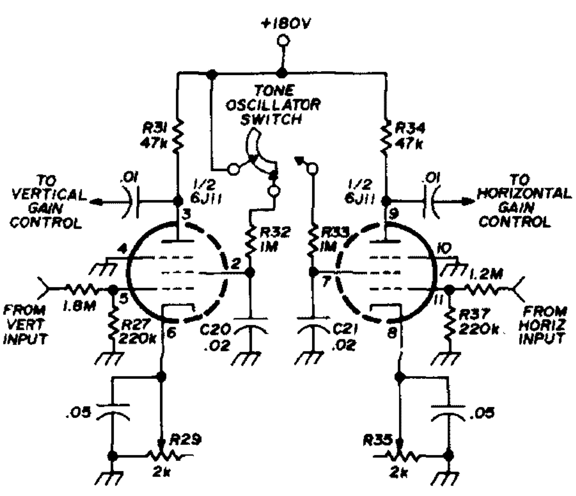


fig. 1. Amplifier for RTTY display uses components included in two-tone test oscillator. All numbered components are original Heath designations. Tube is a 6J11 dual pentode.

One of the objectives of this modification was to minimize the changes to the unit. Wherever possible the original components have been retained. The plate networks were removed as were all components wired to the grid and cathode of each of the two tube sections. The potentiometers which were used to set the tone levels are used to set the cathode bias. The only components added were four resistors (grid) and four capacitors (cathode bypass and plate coupling). The original wiring of the tone oscillator switch remains as a method of disabling the RTTY display when observing the transmitter. The single-tone position with horizontal sweep turned on gives somewhat of a time display at no additional cost.

Fig. 1 shows the circuit as it would appear in the HO-10. Numbered components are original components. Component numbering and values will differ somewhat in the SB-610, but the circuits are similar. Save the few parts removed and it will be possible to return the unit to its original condition with a few minutes of your time and no expense.

Robert Clark, K9HVW

adding carriage return to the automatic line-feed generator

The automatic line-feed generator featured in the January, 1973, issue of *ham radio* can be modified to provide the requirement that the printer be modified with a non-overline kit. It has the advantage that it can be used with model 15s, 19s, Mites, Kleinschmidts and other printers provided that a suitable mechanical mounting can be devised for the microswitch and that the interfacing of the

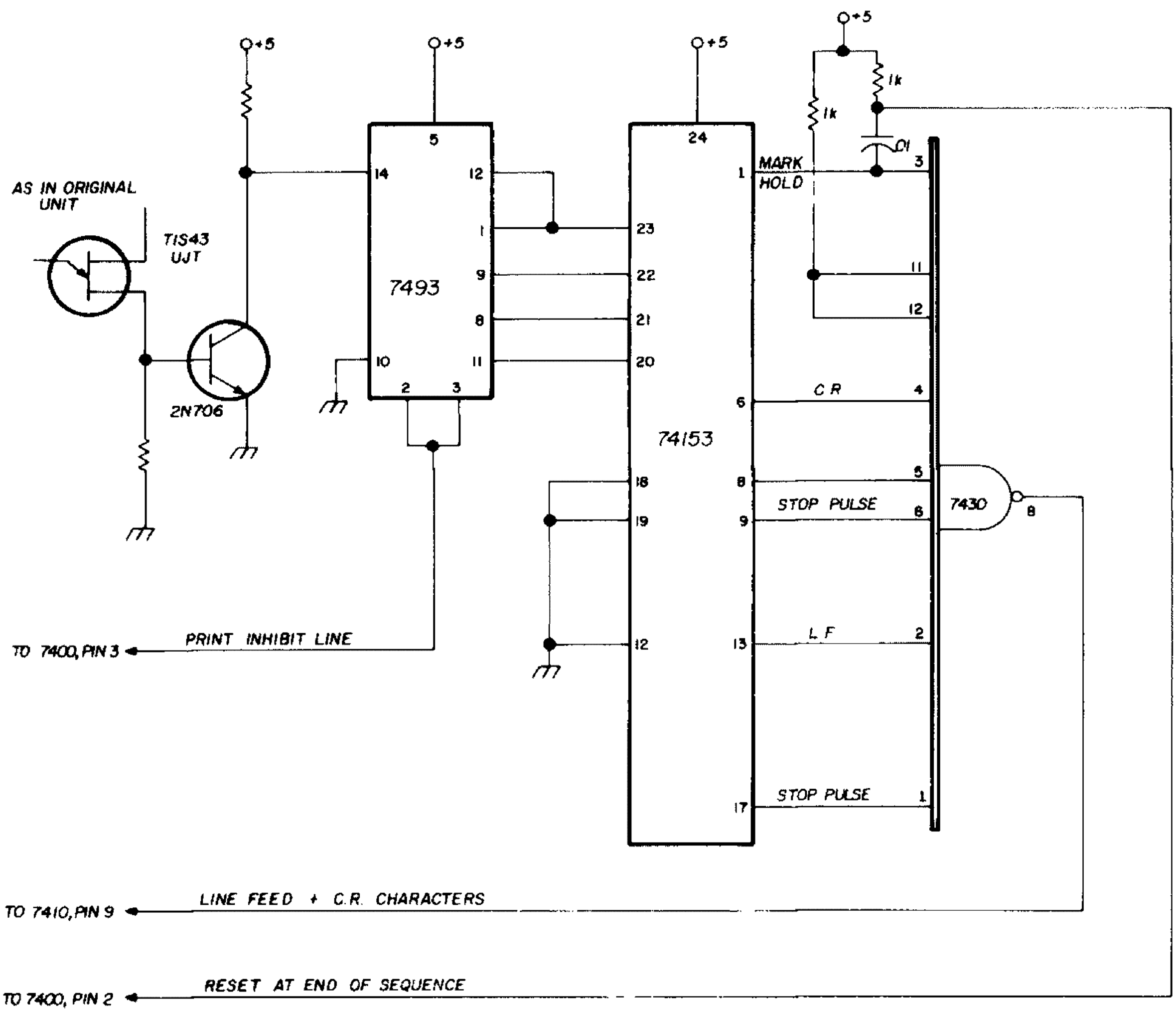


fig. 2. Simple fnodification for adding carriage return to the automatic line-feed generator.

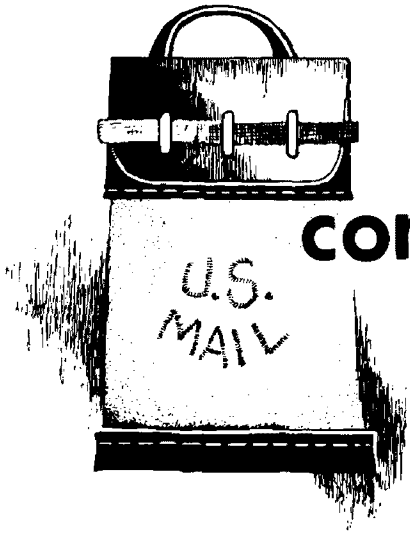
carriage return signal in addition to the line-feed character. As the article explains, the digital device is actuated by a microswitch mounted in the printer so that it is tripped when the carriage approaches the end of a line. The logic then generates the electrical signals for carriage return and line-feed which are inserted in the printer loop, actuate the mechanism, and the printer then returns to normal print. The prevents overstrikes at the right margin.

The carriage return modification does not add to the cost and eliminates the logic board can be made with existing equipment in the station. This is provided for in the TTL autostart unit described in the June, 1973, issue of *ham radio.*

The modification envolves substituting a 7493 and a 74154 for the 7490 and 7442 in the original unit, as shown in fig. 2. A plated epoxy board is available for $7.00 postpaid in the USA.*

Bert Kelley, K4EEU

*Order from the author, Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609.



comments

speech clipping

Dear HR:

I would like to make the following comments in response to W6VFR's letter in the August, 1973 , issue of *ham radio.* If the i-f response of a receiver is measured with the agc system operating, any peaks will appear to be flattened out, but this, unfortunately, does not make it sound any better. The same applies to the case of clipping, which also produces the appearance of a flat response curve without altering the relative strength of the various components of the signal present at any given moment.

The apparent reduction of passband ripple, when the response is measured in the usual way, is therefore of no relevance. The passband ripple of all filters, at both the transmitting and receiving ends of the circuit, are fully additive in their effects.

Choice of filter bandwidth is a matter for compromise which, in the case of standard filters, embodies a lot of practical experience. What is optimum for a single filter must necessarily be non-optimum for a filter to be used in cascade, although results may still be acceptable. To establish that there is "no degradation," however, would require carefully controlled tests covering a wide range of conditions.

It seems important to me that the possibility of using standard filters should not divert attention from the need for filters designed for the job, with the object of producing optimum performance at minimum cost. I feel that interest in rf clipping has now reached the point where the excuse of insufficient demand no longer applies.

Les Moxon, G6XN
Petersfield, England

Dear HR:

I can never understand why so many amateurs knock themselves out designing speech processors, clippers, etc. They really don't make that much difference *on the air* (I speak from experience). All they do is increase background noise and generally cause added and unnecessary confusion.

When I was using a National NCX1000, which had a pretty good rf speech processor, I ran many on-the-air tests. I called DX stations in pile-ups with and without the speech processor enabled, and I'm here to tell you that the speech processor didn't add one damn iota to the score of stations contacted.

I think speech processors are gadgets for guys who are bored and like to fool around with something new. When it comes down to the nitty gritty, the signal that is clean, loud, and in the clear is the one that gets results. A speech processor just eats up extra power. Maybe the speech processor gives the operator a psychological lift, and that's probably okay.

Anyway, the main thing in working hard-to-get stations is a matter of good operating practice, and all the electronic gadgets in the world aren't going to help the idiot-type operator who doesn't have *chutzpah!*

Alf Wilson, W6NIF
Encinitas, California

mexican repeater

Dear HR:

I would like to advise your readers of our new two-meter fm repeater in operation in the Mexico City area. The repeater has been in operation since last August and is the first fully automatic amateur repeater in Mexico. The repeater is sponsored by our club, "Asociación VHF de la Ciudad de México" (Mexico City VHF Association) and is maintained by dues of our members. The call is XE1VHF and frequency is 16/76. Although the repeater is for use by club members, all visitors are welcome. We are also planning a second two-meter repeater as well as a 450-MHz uhf repeater using the call XE1UHF.

Although there is no formal reciprocity agreement between the United States and Mexico, visiting amateurs can contact me at my home address below and perhaps a temporary permit can be arranged. In any event, bring a small rig or a walkie-talkie as a way can undoubtedly be found for a visitor to use the repeater.

Robert N. Green, XE1WS/W2GFO
Palmas 1460
Mexico City 10, Mexico
Telephone: 520-79-93

voltage regulators

Dear HR:

The popularity of the three-terminal voltage regulators (including the Fairchild 7800 series, the Motorola MC7800 series and the National Semiconductor LM 340-T series) has been increasing very rapidly as indicated in several recent *ham radio* and *QST* articles. This popularity manifests itself not only in the difficulty in obtaining these devices because of the heavy industrial demand, but also in these circuits having "unexplained" failures.

In the majority of these failures, the problem can be traced to the lack of non-inductive capacitors being placed at the device's input and output treminals. Because some of the integrated transistors have a f_T of several hundred megahertz, the failure mode is that the regulator breaks into a high-frequency oscillation, resulting in overheating and ultimate device destruction in a matter of seconds.

The part that can cause great consternation for the constructor is that the unit will perform perfectly for hours, days or weeks before the fatal chain of events is initiated. The use of 0.1-μF ceramic-disc capacitors at the input and output pins of the IC is sufficient to stabilize the system.

While the loss of one or two of these regulator chips is not in itself expensive, the resulting destruction of a whole project's complement of TTL or other ICs as a result of being subjected to voltage transients is down right disconcerting! The price of a couple of ceramic discs is indeed inexpensive insurance.

John Perhay, WAØDGW
Savage, Minnesota

aluminum antenna resistance

Dear HR:

I have been on vhf in the Chicago area since 1950, and find that dc antenna resistance is often totally disregarded—cutting signals as much as 50%. I use 300-ohm feedline to 50-, 144- and 220-MHz antennas and can check the dc resistance of line and folded dipole from the shack. It should be only a few ohms. After a windstorm this resistance may actually be more than a thousand ohms due to loss at joints where aluminum is used.

Aluminum power lines used by public-service companies have a special zu-zats put into these joints to avoid high resistance. This zu-zats should be used on all antenna joints where one or both sides are aluminum. If you use 300-ohm feedline and a folded dipole, test the dc resistance. It can be reduced by flashing through—use about 100 volts dc from a small transformer and diode with a 10-μF capacitor in parallel. Flash through the feedline several times. This procedure apparently welds the joints together and, at times, has increased the output signal by as much as 50%.

Ben Hall, W9OVL
Hammond, Indiana

Drake R4 frequency synthesizer

Dear HR:

Ever since I completed the design of the frequency synthesizer for the Drake R-4 receiver (*ham radio*, August, 1972, page 6), I have been looking for a means by which the spurious 10-kHz sidebands might be reduced.

Over the past two years a great deal of additional literature has been published covering phase-locked loops, and that which has come to my attention has been read with considerable interest. Despite all of this information, there appear to be only two basic methods of reducing the sidebands: raise the reference frequency or improve the loop filter (or both).

After reviewing the options, I came to the conclusion that raising the reference frequency would reduce the overall complexity of the synthesizer, at some increase in cost, while an improved loop filter was cheaper but more complicated. I therefore chose the simpler but slightly more expensive approach. A redesigned unit has been constructed and is now in use. The spurious sidebands are now displaced 100 kHz either side of the desired frequency, and are attenuated to an extent which makes them virtually inaudible.

Because of the many letters I have received indicating interest in this synthesizer, I will make available a set of schematics and brief supplementary notes to anyone submitting a self-addressed, stamped No. 10 envelope (4-1/8 x 9-1/2 inches) and $0.50 to cover the reproduction costs.

Robert S. Stein, W6NBI
1849 Middleton Avenue
Los Altos, California 94022

windom antenna

Dear HR:

I noted W4VUO's article in the January, 1974, issue on the four-band, high frequency Windom antenna with considerable interest since I have used that antenna more than any other since Windom first wrote about it. I have used it with great satisfaction on all bands from 10 to 80 meters, inclusive, except the 15-meter band.

My present antenna, used some ten years, is cut a bit longer than indicated by W4VUO's mathematics, 135 feet, 5 inches (41.28 meters) with the feeder attached 19 feet (5.8 meters) from the center. I do not fault the author's mathematics—I derived my length by cut-and-try, and have found the length anything but critical. In fact, in one location I had to drop the ends of the same wire eight feet (2.4 meters) due to the short space available. I noted no loss of efficiency in so doing.

However, the author got out of the band entirely when he called a two-wire feeder, off-center-fed Hertz a *Windom*. The Windom is only a single-wire fed antenna, and reference to the ARRL *Antenna Book* (9th Edition, 1960, bottom of page 191), confirms that statement. It calls the twin-lead, off-center-fed antenna a "miscalled Windom." It further states, ". . . and probably in many cases the line acts more like a single-wire feeder than a parallel conductor one." I would expect it to be a splendid antenna, too, possibly as good as the single-wire fed version.

John E. Waters, DDS, W6EC
Hemet, California

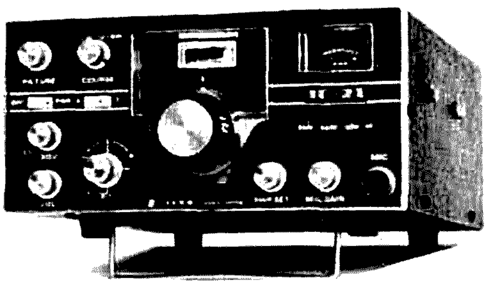
rf clipper

Dear HR:

In all fairness to potential users of the rf clipper for the Yaesu FT-101, Mark 2, described in the *new products* section of the July issue, I should point out that the unit does not work without *any* modifications to the FT-101. Although no modifications are required to any of the circuit boards, the unit does require that a few leads be re-routed via the spare pins on the vfo socket. The modifications can, however, be done in such a way that they are easily reversible, so users do not harm the resale value of their equipment.

Harry Leeming, G3LLL
Holdings, Ltd.
Blackburn, England

new IC-21A from icom



ICOM's great IC-21 base station is now even greater. Now the IC-21A features front-panel control of power level, RIT, plus rf in/out and vswr metering, a discriminator zero-center meter and built in 117 Vac and 12 Vdc supplies. All this in a compact (111 mm H x 230 mm W x 260mm D) 24-channel (22 channel plus two priority) base-station radio including provision for the forthcoming digital vfo.

IC-21A price class $450. Available September, 1974.

antech introduces wide-spaced element 20-meter beam

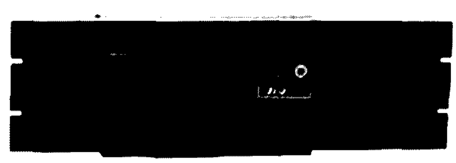
The Antech 20-4 features the same rugged construction and maximum performance offered by the Antech 20-3, and is offered at a special introductory price of $139.95 during September instead of the regular $149.95. The new Antech 20-4 offers 10-dB forward gain, 20-dB front-to-back radio and flat vswr across the 20-meter band. To order the Antech 20-4 or to obtain complete antenna line specifications and prices, write Antech Labs, 8144 Big Bend, St. Louis, Missouri 63119. Bank Americard or Master Charge accepted.

Hy-Gain 18AVT/WB vertical antenna gives true wide-band performance

For omnidirectional performance and 80-meter capability, the Hy-Gain 18AVT/WB is unrivaled. Wide-band coverage, superior construction, brilliant performance and reasonable price make this 80-10 meter vertical a top buy. Automatic switching and true 1/4-wave resonance on all 5 bands. Three traps with large coils for exceptional L/C ratio, high Q. Extremely low radiation angle. 1kW CW, 2 kW PEP. 52 ohms. No. 386.

Hy-Gain Electronics Corporation, 8601 Northeast Highway Six, Lincoln, Nebraska 68507.

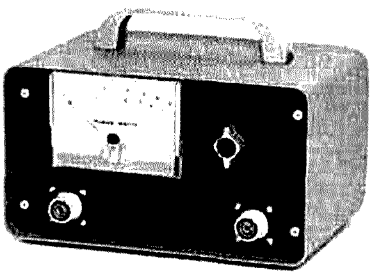
TPL repeater amplifiers



All TPL amplifiers can be ordered in a repeater configuration, designed to mount in a standard 19" (48.25 cm) rack, and is 5½" (14cm) high. The amplifier is rf shielded and is equipped with BNC input and type-N output connectors. It may be ordered with or without power supply and carrier operated forced-air cooling. The extruded rack mounting panel must be specified when ordering, as it cannot be retrofitted to an existing TPL mobile amplifier. TPL Communications, Inc., 13125 Yukon Avenue, Hawthorne, California 90250.

More Details? CHECK–OFF Page 126

new bidirectional wattmeter from dycomm



The model 34 bidirectional wattmeter reads both forward and reflected rf power. It can be left in the line for continuous monitoring and does not require calibration with varying power levels. The taut-band meter movement provides ±3 percent accuracy over four power ranges: 0-10, 0-50, 0-100 and 0-500 W. Bandwidth, defined as the frequency range over which ±3 dB accuracy is maintained, is ±15 percent of center frequency. Any center frequency can be specified from 30 MHz to 470 MHz. Price, $59.95.

ESE moves to larger quarters

ESE, manufacturer of digital clocks, timers, frequency counters and multimeters, is moving to larger quarters in September. Look for ESE at 505 1/2 N. Centinela Avenue, Inglewood, California 90302. The move is brought on, in part, by the enthusiastic response of hams to their digital kits.

The ES 220 and ES 221 are 40-MHz frequency counters; the ES 112 is a 12-hour clock, and the ES 124 is a 24-hour clock; the ES 210 is a 5-range multimeter. All are in stock, ready to ship.

ESE has introduced a *special offer*, consisting of a printed-circuit board and six 0.6" displays, along with instructions, so that the ham who has a drawer-full of ICs can build a 6-digit clock for much less than the $46.95 ESE charges for a complete clock kit.

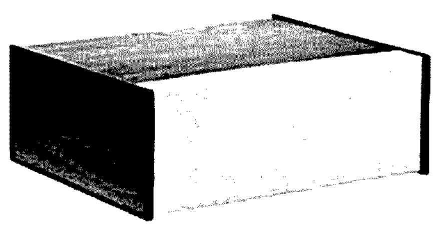
swan's world of amateur radio

"The World of Amateur Radio" is a new 20-page catalog from Swan Electronics presenting the company's line of transceivers, linear amplifiers, fixed and mobile antennas and compatible accessories for the amateur.

Highlighted are three transceivers: the Champion with 700 watts of PEP input on ssb; the 5-band portable Cygnet de novo with 300 watts PEP; and Monobanders in a choice of 75 or 160 watts for mobile 40- or 80-meter operation. Other equipment includes completely solid-state, 5-band transceivers and a 2000-watt linear amplifier.

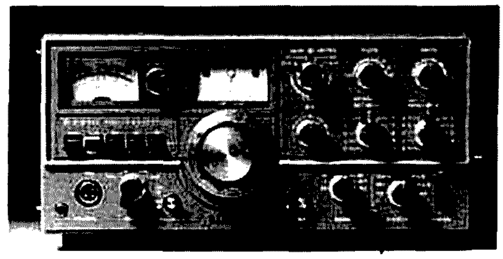
Described are 2- and 6-meter transceivers, a deluxe 600-watt station and several commercial ssb transceivers for two-way requirements. Finally, a line of fixed and mobile antennas are shown along with other accessories. Copies may be obtained from Swan Electronics, a subsidiary of Cubic Corporation, 305 Airport Road, Oceanside, California 92054.

equipment enclosures



The distinctive, smart looking enclosures used to house the Ten-Tec line of Amateur Radio equipment are available to hobbyists and home-brewers, as well as to other manufacturers. Forty standard models are available in four basic series and two color schemes. Give your construction project a professional look with a Ten-Tec enclosure. Available through your authorized dealer or directly from us. Write for descriptive brochure. Ten-Tec, Inc., Sevierville, Tennessee 37862.

the popular Kenwood TS-520



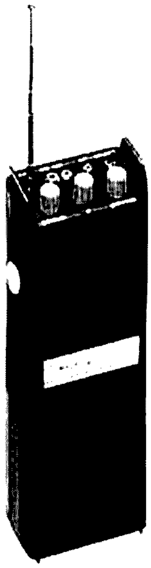
Never has a transceiver been received with such overwhelming enthusiasm as has the new Kenwood TS-520. It is a superbly built 5-band transceiver for ssb or CW operation, boasting built-in ac and dc power supplies, built-in vox with adjustable gain, delay and anti-vox, 1-kHz dial readout, ultra stable fet linear vfo, built-in noise blanker, built-in RIT circuit and RIT indicator light . . . and much more. $629.00 at selected Kenwood dealers.

electronics bench manual

The *Electronics Bench Manual* is a massive collection of data commonly required during the bench phases of electronic experimentation and construction. This device-oriented document contains sections on electronic components (semiconductors, electron tubes, electrostatic devices) as well as mechanical movements and actuators, hardware, finishes and housings, application diagrams, construction techniques, bench layout and support facilities, and much more. A high-density document with large-page format packed at an equivalent of about 2000 words per page; the semiconductor section itself contains almost 100,000 words of description. Each section of the manual is an independent booklet, removable for use at the bench. The entire publication is mounted in a rugged, practical looseleaf binder. 432 pages, 823 illustrations, 86 tables; priced at $17.95 (ppd/U.S.) from Technical Documentation (Tdoc), Box 340, Centreville, Virginia 22020.

More Details? CHECK–OFF Page 126

regency offers two-meter hand-held radio



A new 5-channel, 2-meter narrow band fm hand-held transceiver is now available from Regency Electronics, Inc., Indianapolis-based manufacturer of a complete line of FM radios.

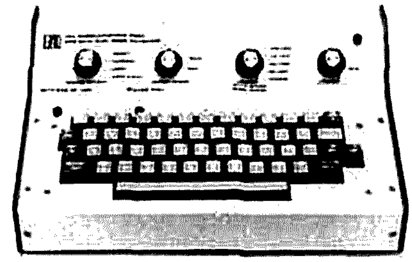
Designated the HRT-2, the radio is the portable and convenient answer to two-way radio needs. A double-conversion, superhetrodyne receiver design employs two ceramic filters. The transmitter and receiver sections both employ bandpass circuitry. Complete operation from 3 simple controls, plus high-low power switch. The radio boasts "American Made quality with Regency reliability." Regency Electronics, Inc., 7707 Records Street, Indianapolis, Indiana 46226.

new comcraft CSP50

The Comcraft Company announces the introduction of a new all solid state, 2-band frequency synthesized fm transceiver. The new transceiver, called the CSP50, features operation on both 2 meters and 1¼ meters with 25 watts output and 5-kHz frequency-synthesized channel steps. Operating modes provided include simplex, split transmit and receive with all of the popular repeater offsets. For further details write the Comcraft Company, P.O. Box 266, Goleta, California 93017.

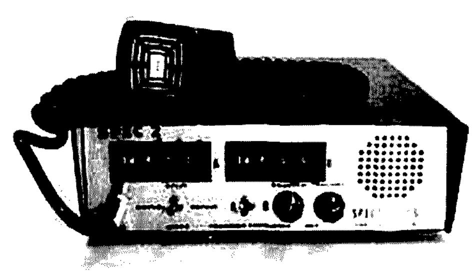
More Details? CHECK—OFF Page 126

dual-mode keyboard



HAL Communications Corp. has introduced the DKB-2010 Dual Mode Keyboard. This keyboard features both Morse and RTTY capability, and also incorporates new features such as N-key rollover, 3-character buffer memory, identifier providing one key for call letters, and two keys which can be programmed for any 3 characters. Optional 64 and 128 character buffers, which can be loaded off line and released when desired, are available. Complete information available from HAL Communications Corp., Box 365H, Urbana, Illinois 61801.

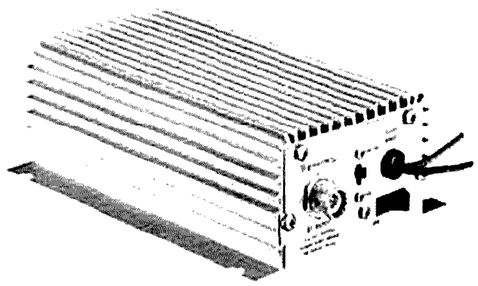
SPEC II synthesized fm transceiver



Now you don't have to wait for full 2-meter coverage in 10-kHz steps, (extended range model also available for MARS, etc.) The SPEC II combines proven Motorola portable transceiver circuitry with the quality RP synthesizer. Output is typically 2 watts with receiver sensitivity typically .25μV across the entire band. For full information contact Spectronics, Inc., 1009 Garfield, Oak Park, Illinois 60304.

More Details? CHECK—OFF Page 126

compact vhf power amps



Though new to amateur circles, Vibratrol is well-known as a major supplier of commercial-service solid-state power amplifiers. Vibratrol's broad line of high- and low-band amplifiers are quite competitive in price, and unusually compact due to their uniquely effective heat sinking.

For high-band – 2 meters – Vibratrol offers six models with outputs to 150W. Three are low-drive models (3 W maximum); the others require 10 W for rated output, though all will operate with only 1 W. Vibratrol, 505 Harvester Court, Wheeling, Illinois 60090. (312) 541-5110.

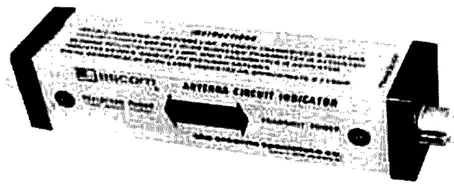
little giant antenna for use in small areas

Stan Byquist's Little Giant antenna is not exactly new – he received his patent on it back in the 1950s and he has recently resurrected the design.

In brief, the Little Giant is a highly compressed single-band antenna that can be used in situations where conventional antennas are out of the question. The largest model, for 80 meters, is only 27-inches (68.6-cm) high and 32-inches (81.3-cm) wide; even smaller models are available for use on 40 through 10 meters. Bandwidth is necessarily small, as would be expected of such a small and, therefore, high-Q antenna. User reports have been quite favorable considering that any drastically shortened antenna is bound to be a compromise in performance. Amateurs with a space problem should contact Stan at Apollo, Vaughnsville, Ohio 45893.

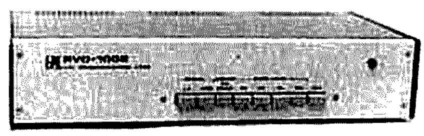
More Details? CHECK–OFF Page 126

ascom LED vswr indicator ASM-104



A product of advanced space age technology, the ASM-104 indicator is a go - no go device for marine and amateur radio systems. The indicator is designed to be left in the line at all times. When the transmitter is keyed, one red LED (light-emitting diode) lights; if an antenna mismatch exists the other LED will light, indicating that the system should be checked. The ASM-104 does not introduce any losses or mismatch to the system. Frequency range is 144 to 225 MHz. Power handling capability is 25 watts. Suggested Resale Price $19.95. Contact your local Antenna Specialists distributor or write: The Antenna Specialists Co., 12435 Euclid Avenue, Cleveland, Ohio 44106.

visual display unit



The HAL RVD-1002 provides a visual display of 5-level (Baudot) code TTY when driven from HAL ST-6 or other terminal unit. It features operation at 60, 66, 75 and 100 wpm, selectable unshift on space, and 1000 character display. Output is standard 75-ohm TV video. Use as a read-only display for ham and commercial RTTY, or combine with the new DKB-2010 for a complete machine replacement. Complete information available from HAL Communications Corp., Box 365, Urbana, Illinois 61801.

More Details? CHECK—OFF Page 126

gregory specializes in fm mobile equipment

Gregory Electronics Corporation has added to its vast inventory the General-Electric Transistorized Progress Line used fm two-way mobile communication equipment. These TPLs are available for use in six- and two-meter frequency bands.

Gregory Electronics is the largest distributor in the world specializing in used fm mobile communication equipment.

Gregory Electronics is interested in purchasing used late model equipment, manufactured by General-Electric, Motorola, and RCA.

Readers of *ham radio* may still write for the current 1974 catalog. A new 1975 catalog will be available in late fall. Gregory Electronics Corporation, 249 Route 46, Saddle Brook, New Jersey 07662.

tri-tek — new or surplus components

Tri-Tek would like to take this opportunity to thank the many *ham radio* subscribers who have become our good friends and customers. We're happy to have been able to help with your past projects and will do our best to continue to offer the finest in new and surplus components and materials at reasonable prices.

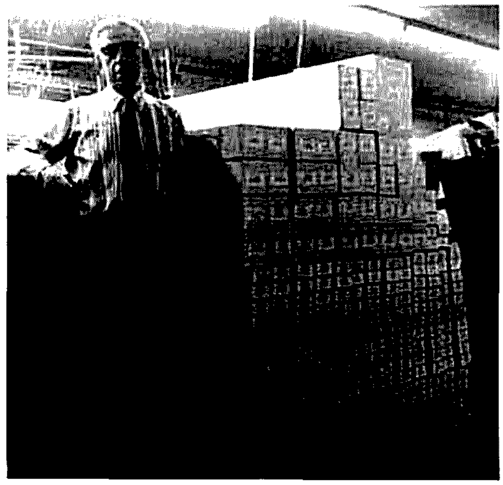
For our latest complete listing, send a 10¢ stamp to Tri-Tek, Inc., P.O. Box 14206, Phoenix, Arizona 85063.

switchable keyer

A front-panel switch in Matric's Model 10A Electronic Keyer permits both self-completing dots and dashes or only self-completing dots. The isolated output reed relay handles up to 100 Vdc at 20 watts, ideal for low-power rigs, particularly with an idle current of 3 mA and operating current of 12 mA.

A perfect companion is the Model 11A Paddle with easily adjusted spacing, a damped lever and weighted base.

we're not big . . .
but we're thorough



Les Cushman (W1AWZ) the man responsible!

Besides saving hams a lot of money, we contribute a good full line of quality amateur radio antennas designed for easy installation. No professional help is needed — just take out of the box and household tools will do the rest simply and quickly. When we develop a basic antenna, we follow through with a complete line to satisfy popular demand. Distributors worldwide stock Cush Craft Antennas.

Thanks for your support and encouragement. Cush Craft, 621 Hayward Street, Manchester, New Hampshire 03103.

new digital dial

Matric's new Model 22 Digital Dial provides a 100-kHz readout to 100 Hz. Modular construction enables the use of an interchangeable input module to match the requirements of most ham gear.

A crystal time base ensures 100-Hz accuracy while a non-blinking display gives five up-date readings per second. Two switched inputs are provided along with a calibrate control for set-up.

A one-wire connection to your vfo buffer or output jack plus either 115 vac or 12 Vdc is all that is required. The input requires 70 mV rms to a 1-megohm, 20-pF circuit.

More Details? CHECK—OFF Page 126

new concept in two-meter antennas

Hy-Gain Electronics announces a new 2-meter gain antenna for the mobile, fixed or marine operator. The antenna, Hy-Gain Model 270, is designed to overcome the difficulties of current gain mobile antennas. It eliminates hard tuning, high swr, whip-flex fade and pattern distortion due to irregular ground plane.

It is optimized for the fm operator with less than 1.5:1 VSWR from 146 to 148 MHz (actual 2:0 vswr bandpass is 144 to 150 MHz). The design develops 6 dBi gain through the use of two 5/8-wave sections with their own decoupling system. By operating the antenna independent of car body ground it can be factory pre-tuned. Also, the Hy-Gain 270 has minimum pattern distortion due to irregular ground plane. Because it is all fiberglass, whip-flex fade is gone.

Additionally, because the Hy-Gain 270 is sealed in fiberglass, it will deliver outstanding performance month after *month. No deterioration of performance* due to corrosion of the antenna or feedpoint.

The Hy-Gain 270 can be used fixed, mobile, marine or even as a repeater antenna. For fixed usage a 271 mast bracket is available. The Hy-Gain 270 is the first gain 2-meter system that can be used anywhere without problem. Ham net is $39.95 for the 270 and $4.95 for the 271.

Madison

Madison Electronics has in stock a wide range of antennas, towers, rotators and other important antenna accessories.

They're featuring the Hy-Gain TH6DXX 6-element tribander these days and suggest that you call or write for a quotation. Madison also has Tri-Ex towers, Belden coax and Ham-II rotators, all at attractive prices.

Also be sure to ask for a free flyer from Madison Electronics Supply, Inc., 1508 McKinney Avenue, Houston, Texas 77002

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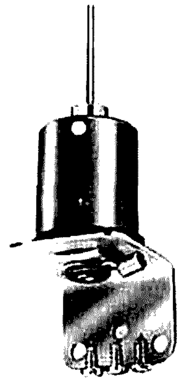
larsen features trunk mount mobile antenna



This Trunk Mount mobile antenna gives a 3-dB gain over a quarter-wave whip. It comes complete with mounting screws, stainless-steel bracket antenna mount, and can be secured complete with coax if desired. Designed for maximum efficiency and minimum wind drag; it will handle a full 150 watts with no difficulty. Requires no coil tuning or special mounting adapters. All Larsen antennas are fully guaranteed. Models are available for both 2 meter and 432 MHz from Larsen Electronics, Inc., 11611 N.E. 50th Avenue, Vancouver, Washington 98665.

new expanded stock

This autumn the fellows at M. Weinschenker will be able to supply components and service for both amateurs and experimenters better than ever before. This summer they have been working to improve their stock both in variety and depth to better give you the backup you want.

Weinschenker has tried hard to keep up with the rapidly changing world of electronics, particularly by supplying quality U.S.-made electronics for your special projects.

A check of their new Catalog number 10 and their latest ads in *ham radio* will show many interesting items; for example, the wide variety and low cost of LED readouts in many colors to help the builder turn out professional looking and performing projects at budget prices.

Each month be sure not to miss the Weinschenker ad in *ham radio*.

More Details? CHECK—OFF Page 126

electro-monitor fm receivers feature unique simo

"Get In On The Action" best describes the new Electro-Monitor FM monitor receivers featuring "SIMO," the only multiple frequency simultaneous fm receivers on the market today. It monitors any two fm frequencies, high or low, at the same time enabling you to hear two public service transmissions simultaneously — or both sides of a duplex communication.

Electro-Monitor receivers come complete with crystals and receive low band, high band or uhf frequencies. Additional information from Electrosonics, 34 E. Logan Street, Philadelphia, Pennsylvania 19144.

BC electronics — a history of serving the amateur

Ben Cohn of BC Electronics was one of the few early dealers in military surplus to cater to hams. Ben started in business as Quad Electrical Supply in a large, cheap, loft space on the northwest side of Chicago. In the early 1950s a move was made to South Michigan Avenue to create a surplus electronics center. Today, BC Electronics, 1249 Rosedale Avenue, is the last surplus dealer in Chicago catering to hams and one of the few remaining original surplus dealers still serving amateurs.

More Details? CHECK—OFF Page 126

SAROC returns to the hotel sahara

It's been a decade since SAROC and Hotel Sahara got together to organize the first ham gathering. In those ten years, SAROC has blossomed from 250 registrants to a record of nearly 1500.

At the same time, Hotel Sahara's grown too! One thing, however, hasn't changed, and that's the personal service and attention you'll receive from our staff.

We're proud to have SAROC back at the Sahara and we intend to show you a good time.

antenna mart offers many product lines

The Antenna Mart, owned and operated by George McKercher, WØMLY, has been located in Pippey, Iowa for eleven years and has over 6,000 square feet of floor space. Products manufactured include the RX1 heavy duty rotator, SW5 remote coax switch and the MX1A plug-in mixer for Collins receivers. Distributor for HyGain, Newtronics, Mosley, KLM, Wilson Antennas, CDR Rotators, Rohn, Heights and Tri-Ex Towers. Specializes in parts, government electronic surplus and custom work. Antenna Mart, Box 7, Rippey, Iowa 50235.

More Details? CHECK—OFF Page 126

electronics center, inc. dallas, texas

You ought to be part of our monthly sidewalk sale, held on the first Saturday of each month. Electronics Center institutionalized this event that brings more than 400 hams and traders to the center of ham activity for the Southwest.

The staff and facilities are operated by Walt, W5ZYA, Al, W5PXH, Larry, WA5BEN, John, K5HIH, Pete, WB5IGL, and Danny, WB5DNT. Come see us and all of the popular hf, vhf and uhf gear at 2929 N. Haskell, Dallas, Texas 75204.

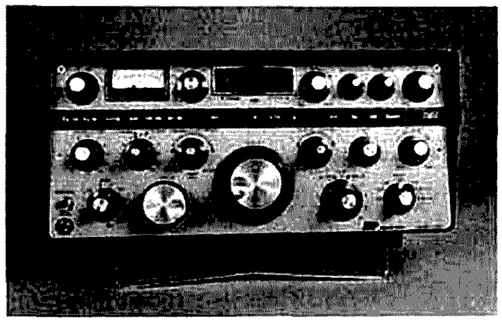
ye olde ham spirit

Whether you are local, from another part of the USA or anywhere in the world, feel free to hang your hat at Barry Electronics, 512 Broadway in New York City. All hams are welcome whenever they're in the big city and an extra special greeting awaits any overseas amateur when he is visiting. Come on in, look around and talk, you'll find a warm reception from Barry, W2LNI, and his trained courteous staff.

Barry has a very unique collection of miscellaneous ham gear that you will not usually find elsewhere. Heavily stocked on tubes, especially Eimac, Barry also has miscellaneous parts for anything. He stocks Millen, Barker & Williamson, and two-meter fm transceivers including the IC-230 and Clegg FM 27B.

Be sure to make Barry's part of your next visit to New York.

CX-11 from signal one



Signal One Corporation announces the new CX-11 completely solid-state integrated station, designed to be the ultimate in desirable features and reliability. It uses gold-plated sockets and pins for easy removal of all transistors, integrated circuits, and circuit boards. New features include: five bandwidths of selectivity, variable notch filter, short-proof power supply, solid-state broadband final that requires no tuning, six-digit LED frequency counter and a new concept in front end design making all other commercial ssb equipment obsolete.

These specifications and other features make the CX-11 unequaled in communications technology.

QRP rig from matric

The new Model 50 CW Transmitter by Matric includes an ac power supply, antenna relay and full break-in keying.

This crystal-controlled, 15-watt input rig will cover 160, 80 and 40 meters. It has plug-in toroid coils and features zener-regulated, chirpless keying and clean T-network rf output.

Also look soon for the Model 40 CW Transceiver, a totally portable 2-watt design.

new keyboard and encoder kit available

Here is a top-quality, fully professional data-entry keyboard at a reasonable price. A 49-key system with internal ASCII encoder and switch debouncer. Shift and control keys are provided along with two user-defined keys. Keyboard uses all

More Details? CHECK–OFF Page 126

brand new keyswitches and a modern IC and diode encoder circuit.

Output is a standard parallel ASCII type code that can be used with almost any type computer system. The circuit board is double sided with plated-through holes for quick, easy—no jumper—construction. Requires +5 and −12 volts dc for operation. Standard TTL logic levels at the output.

For terminals, calculators, TV typewriters, RTTY displays, video titlers, teaching aids. KBD-1 Kit . . . $39.95 postpaid from Southwest Technical Products, 219 Rhapsody St., San Antonio, Texas 78216.

versatile circuits ease system design

Dodd Digital Design announces the "Series 100" line of etched-circuit cards, which provide the amateur with state-of-the-art logic circuits. Each of the 14 cards performs one complete function; several cards may be interconnected to construct equipment such as an automatic logging system, a CW/RTTY generator, a large memory, etc.

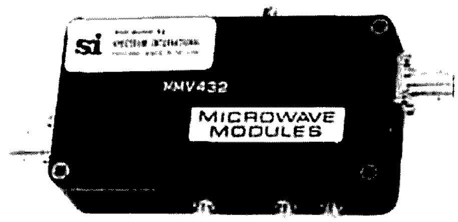
The plug-in cards are of professional-quality construction. For a catalog, send 50¢ to Dodd Digital Design, 234 Waples Park, Fairfax, Virginia 22030.

new wide band yagi from KLM

KLM introduces the 140- 150- 12C circularly-polarized wide band Yagi. While primarily designed to work with the new OSCAR 7 super satellite, this antenna offers flexibility of 2-meter operation that is hard to match. Exceptional performance is offered whether the stations on the end are horizontal or vertically polarized. The antenna features 12 elements on a seven-foot boom, complete with phasing harness, balun, and all stainless steel hardware. Options are a circular 6-element 432-MHz "add on" kit for the complete OSCAR system on one boom.

More Details? CHECK—OFF Page 126

uhf varactor multiplier triplers



The easy way to get on the "next" U.H.F. band. From 2 metres, the MMv432 typically delivers 14W for 20W drive. Similarly the MMv1296 gets you from 432 to 1296; 12W typical output for 20W drive. These wideband units cover 420 to 460 MHz without returning. No dc power supply required, just plug in the drive and connect the antenna.

From Spectrum International, P.O. Box 1084, Concord, Massachusetts 01742.

versatile speech processor

The Model 60 Speech Processor by Matric increases the talk power of your station by increasing the average-to-peak power ratio of your speech.

Featuring instantaneous attack and release with a process threshold of 1.5 mV rms (Hi-Z) and 400 =V rms (Lo-Z) the Model 60 offers a frequency response of 300-3000 Hz ±1.5 dB at a maximum output impedance of 2500 ohms.

In/out switch and gain control for ease of operation. Matric, RD #1 Pone Lane, Box 185A. Franklin, Pennsylvania 16323.

collins expands amateur operation

Collins Radio is placing increased emphasis on amateur activities with formation of a new amateur radio business operations group. Manager of the new operation is Joe H. Beler, W5WY/Ø, with Collins 14 years in various management positions. A ham more than 30 years, he has an extra-class license. Formerly a Colonel in the Air Force, Beler was re-

More Details? CHECK–OFF Page 126

sponsible for introducing operational hf ssb to the Strategic Air Command.

Jerry Carter, WAØZRW, a ham 11 years, is in charge of amateur sales. His business phone is 319-395-4507.

Arnold Verdow, WØLIJ, longtime Collins employee and active ham, joins the new operation, responsible for product support. His phone is 391-395-3393.

The amateur radio business operation is part of Collins' Telecommunication Equipment Division, Cedar Rapids, Iowa. Collins is a Group of Rockwell International Corporation.

G&G offers free catalog of military surplus

G&G Radio Electronics Company at 45-47 Warren Street, New York, New York has served the electronics industry since 1920. Their free catalog describes a tremendous inventory of electronic surplus material of both WW II and Vietnam vintage.

Write G&G for anything you need not listed in their catalog. They will do their best to get it for you at lowest current prices. Inquiries from quantity buyers are invited. See G&G's ad in this issue of *ham radio*.

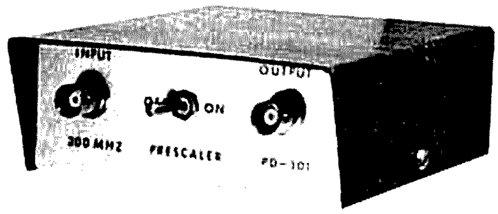
erickson communications serves chicago hams

Catering only to radio amateurs, Erickson Communications offers over-the-counter service to hams in greater Chicago while also serving the mail-order field. Erickson maintains a good stock of vhf uhf equipment and accessories at all times, and since owner Jim O'Connell (W9JZK, ex-KR6NR) also works the low bands, Erickson covers that market as well.

Erickson services and installs all lines they handle. Erickson is convenient to Eden's Expressway in north suburban Skokie. Erickson Communications, 3501 West Jarvis, Skokie, Illinois 60076. (312) 677-2161.

More Details? CHECK—OFF Page 126

300-MHz prescaler



Model PD301 is a 300-MHz prescaler designed to extend the range of your counter 10 times. This prescaler has a built-in preamp with a sensitivity of better than 50 mV at 150 MHz, 100 mV at 260 MHz and 175 mV at 300 MHz.

The prescaler is available with or without a self-contained power supply. The PD301 has been successfully tested with many popular counters. The PD301 is available from K-Enterprises, 1401 North Tucker, Shawnee, Oklahoma 74801.

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Superfilter — modern crystal filter theory yields superior performance at a remarkably low price. Bandwidth of 2150 Hz at 6 dB is ideal for ssb. Rejection of 50 dB at ±300 Hz and 90 dB at ± 700 Hz from the passband edges, suppresses even the strongest interference. What's more, 90-dB rejection is maintained down to 100 Hz and up to 15 MHz leading to simple i-f designs. See our Superfilter ad in this issue for more information — Signal Management Sciences.

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Heights Manufacturing Company offers special Screw-Operated Hinged Bases and Fold-Over Kits for operation with their famous line of aluminum towers.

The Screw-Operated Hinged Base (SOHB) gives tremendous leverage to an individual working on the tower plus great safety in the inherent self-braking of screw type actuation.

More Details? CHECK—OFF Page 126

For easier and faster operation, Heights also offers winch operation by the substitution of a winch, cable and pulley system. In addition the SOHBs described above can be converted to this type of operation by the installation of these new parts.

Get complete information, not only on these special SOHB and Fold-Over Kits, but also on their complete line of "Standard" aluminum towers from Heights Manufacturing Co., Almont Heights Industrial Park, 4516 North Van Dyke Highway, Almont, Michigan 48003.

midwest's ham headquarters

Chuck Schecter, W8UCG, welcomes you to call Electronic Distributors, Inc., (616) 726-3196 or TELEX 22-8411 for quotes on towers, antennas, electronic and ham gear (both new and used), accessories, etc. For over 36 years a leader in the electronics industry as communications specialists. Your one-stop electronics supply center! Daily from 8:30 - 5:30 – Saturday from 9 - 4. Satisfaction guaranteed. All the best name brands at Electronic Distributors, Inc., 1960 Peck Street, Muskegon, Michigan 49441.

flexible two-meter antennas

Hy-Gain Electronics announces two new "rubber-ducky" antennas specifically tuned for the ham. The new Model 274 is a plastic-coated helical flexible antenna which terminates in a male BNC fitting and will fit directly on the many hand-held units employing BNC output connectors such as the Wilson, Clegg, VHF Engineering and other popular units. Ham net for the Model 274 is $9.00.

The new Model 275 is terminated in a metal shell PL-259 fitting for two-meter units which previously could not use a flexible antenna. It is tuned specifically for the Drake TR series and similar portable units. Ham net for the Model 275 is $7.00.

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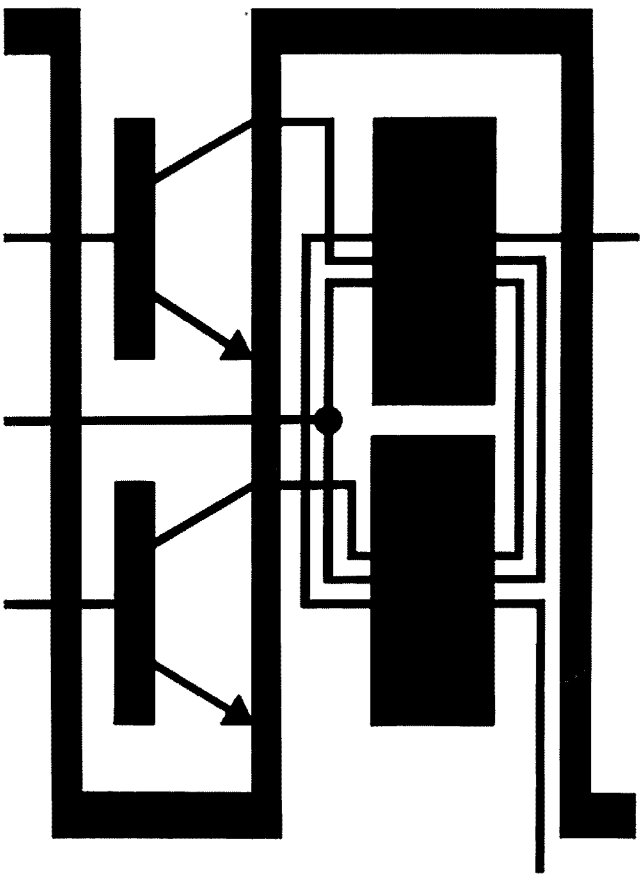
this month



high-efficiency rf power amplifiers

October, 1974
volume 7, number 10

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Hilda M. Wetherbee
assistant publisher
advertising manager

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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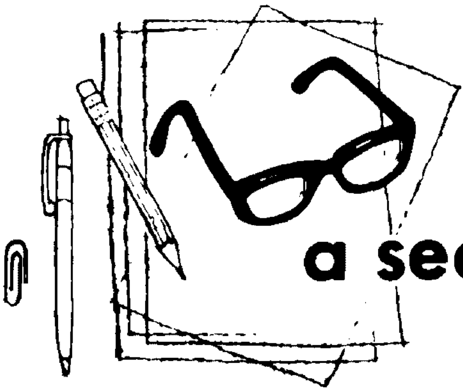


a second look

by jim fisk

In our ever more populated world, with its growing proliferation of electronic gadgets, it's the rare amateur who hasn't been bothered at one time or another by interference complaints. In the 1930s, as more and more broadcast stations came on the air, it was BCI. Then, in the late 1940s, it was TVI. Now, with exotic solid-state stereo and quadraphonic systems, the interference problem has become more widespread. It's also much more difficult to cure. Nor, as I pointed out in this column over a year ago, is the problem limited to amateurs. Consumers who live near high-powered television and fm broadcast stations regularly complain of unwanted interference. And the 900,000 or so class-D CB stations receive a share of the blame as well.

Clearly, the problem can be effectively solved only by proper design and construction at the manufacturing level. Why, you may ask, isn't this done now? Wouldn't it be a lot easier to properly design the equipment in the first place than to try to cure the problem piecemeal after the equipment is installed in the consumer's home?

The answer lies in the economics of the design and sale of equipment in a highly competitive market. The manufacturers are obviously reluctant to add filtering or lead bypassing that would increase the size of the price tag. Until recently, in fact, they have contended that only 1% of home entertainment equipment operates in a strong rf environment which requires additional lead filtering or bypassing. However, an Electronics Industries Association spokesman acknowledged recently that the widespread growth of two-way radio systems, as well as higher-power a-m and fm broadcasting stations warrants another look at the manufacturers' present position. However, with dwindling profit margins brought on by inflation, I see little chance manufacturers will voluntarily do anything to solve the RFI problem.

What is needed is Congressional or FCC action to require all manufacturers of TV sets, stereo systems, and a-m and fm receivers to build interference suppression into their designs. Although the late Rep. Charles Teague of California introduced Congressional legislation in 1973 which would have required radio and television sets to meet FCC standards for filtering out interference, that bill, which is still pending, did nothing about radio interference to audio equipment. And, according to Rep. Torbert Macdonald of Massachusetts, Chairman of the House Commerce Subcommittee on Power and Communications, there is considerable legislative opinion that the FCC can require the manufacturers to put lowpass filters into TV receivers without additional legislation.

The ARRL Radio Frequency Interference Task Group, which has been working on the problem for over a year, has put together a packet of material which may prove helpful if you are experiencing rf interference problems. The packet, which includes a wealth of useful data, is available by sending a large (9x12-inch), self-addressed Manila envelope with $0.40 postage to Ted Cohen, W4UMF, 8603 Conover Place, Alexandria, Virginia 22308. Included is a questionnaire which will assist the ARRL RFI Task Force in its work with the FCC and representatives of the electronics industry. Let's all get behind this worthwhile effort.

Jim Fisk, W1DTY
editor-in-chief

presstop

AMATEUR RESPONSE TO CURRENT FCC DOCKETS DISAPPOINTING as few amateurs take time to comment on the many important FCC actions now pending. Let's get with it!

Three Long-Awaited Repeater Actions (repeater linking, cross-banding and automatic control) have received a grand total of two comments thus far, both on automatic control (one wants rules tightened, other wants no rules at all!).

RACES Docket, on the other hand, has had lots of response -- 95% of it from RACES people, and almost 100% of that "anti-ham!" Comments range from "RACES needs paid professional operators -- hams are not competent emergency communicators" to one state organization that proposes taking the present RACES band segment away from the amateur service and reassigning it to state and local governments for full-time RACES use! One thing is certain: RACES has not received good amateur support in most areas.

REDUCED FEES FOR AMATEUR LICENSES offered in Docket 19658 issued August 12. License modification without renewal would be reduced to $5, renewal or new licenses to $6. Special calls remain at $25. Comments by September 20, replies by October 4.

SHARED INDUSTRIAL USE OF 420-450 BAND PROPOSED in FCC Docket 20147, adopted August 21. Docket would permit operation of HIRAN, a high-accuracy radio-location system developed for precise location of offshore oil-drilling rigs. System would use frequencies within the amateur/government-radio-location band on a non-interfering temporary basis through January 1, 1978. Comments are due by November 4.

ARRL ADVISORY COMMITTEE OPENINGS on the Contest, DX and VHF Repeater committees upcoming as many present appointments expire January 1, 1975. Nominations to fill the open slots are solicited, and the proper forms for making nominations are available on request from ARRL headquarters in Newington, Connecticut.

HAWAII SITE FOR NEW 1975 HAMFEST. Promoted by Leonard Norman of SAROC/Las Vegas fame, this new show is in addition to SAROC. A week-long affair is planned with package deals from all major U.S. cities. For developing details write to Leonard at SAROC, Box 945, Boulder City, Nevada 89005.

NATIONAL RADIO RIDES AGAIN. All assets of the former National Radio Company have been purchased by new National Radio Company, Inc., with plans to resume manufacture of most of the former company's product lines in Melrose, Mass.

PROBLEM WITH KEYBOARD CW BUFFS is proper IDing. FCC monitors are not set up to copy super-high-speed CW, so perhaps FCC rule limiting RTTY CW ID to 20 wpm should be extended to all modes. Forearmed is forewarned...

FCC COULD LICENSE ALIENS under provisions of Senate bill S.2457 now under consideration. This bill would enable many now prohibited to become licensed, and expedite licensing of others now eligible by bypassing considerable red tape.

ILLINOIS STATE POLICE APPLY FOR 2000 CB LICENSES, put FCC on spot as to whether this is "proper" CB use. Police/citizen groups are using CB in several areas, but question seems to be what impact CB licensing of a large force would have.

AT TEXAS VHF-FM SOCIETY CONVENTION in August FCC Amateur and CB Division Chief Prose Walker delivered a Special Temporary Authority for activation of TIRS (Texas Intercity Relay System) that permitted demonstration of the capabilities of this sophisticated multi-repeater linkup.

Prose also discussed restructuring of amateur radio, and reviewed a variety of possible approaches including a separate license for HF or VHF/UHF (with distinctive prefix -- AA-AL block is already ours, and attempts are being made to break N and NA-NZ prefixes loose from the Navy).

ARRL NEW ENGLAND DIVISION DIRECTOR CHAPMAN RETIRES effective December 31, but Bob, W1QV, plans to continue as President of the ARRL Foundation until the end of his term next March.

high-efficiency rf power amplifiers

Design and construction of solid-state rf power amplifiers that offer operating efficiencies of 90 percent or more

Frederick H. Raab, WB8LQK, 11762 Cedarcreek Drive, Cincinnati, Ohio 45240

Since the demise of the spark gap, radio amateurs have been plagued by inefficient transmitters. The transistor eliminated the power consumed by the filaments of vacuum tubes, but the inefficiency inherent in class-A, -B and -C amplifiers remained. Now class-D techniques (also called switched-mode or class-S) offer a means of eliminating some of the remaining inefficiency. These techniques can be applied to both audio and rf amplifiers, as well as to voltage regulators.

There are a lot of reasons an amateur might want to use class-D, including less input power for the same output, more output for the same input, smaller power supplies or batteries (lower energy consumption), and smaller heatsinks and transistors. In addition, the transistors don't have to be linear.

You might use class-D to build a transmitter with a legal *output* of 900 watts. High-efficiency operation will keep this rig from heating your operating room at the same time. Class-D operation not only reduces your electric bill, it result in saving even more power at the generat ing plant (1 to 2 watts there for each wat saved in your transmitter).

Why are ordinary amplifiers ineffi cient, anyway? When used as a class-A, -B or -C amplifier, a transistor (or tube, fet etc.) sustains a non-zero voltage while it is conducting current. Whenever both volt age and current are present, power is consumed. The amount of power con sumed depends on the class of operation — what portion of the time the active device is actually conducting cur rent.

Class-A amplifiers conduct current al the time, and when amplifying a sine wave, efficiency is no greater than 50 percent. Since class-B amplifiers conduct current half the time, they can achieve 78.5% efficiency with sine-wave signals. Class-C amplifiers conduct current less than half of the time and do not, theoretically, have limited efficiency. However, practical limits (drive power, peak current, etc.) make it difficult to achieve very high efficiencies; maximum typical efficiency for class-C operation is 85 percent.

Class-D operation can provide more efficiency because it avoids the conditions of simultaneous non-zero voltage and non-zero current which cause power to be consumed in the transistor. The transistors act as switches. When the transistors are *off* (open), the current is zero, so no power can be dissipated. When the transistors are *on* (closed), the transistors have (nearly) zero voltage drop across them, so again no power is dissipated. So long as the transistors can be switched

Author Raab, who was previously licensed as WAØATT, is employed by the Cincinnati Electronics Corporation in Cincinnati, Ohio, manufacturers of electronics equipment for the government.

fast enough and saturated well enough, high-efficiency operation is possible.

The difference between linear and switched-mode amplifiers is best shown by an example. Suppose that you have a lamp which consumes 100 watts when connected to 100-volts dc, and that you want to dim it to 50 watts (fig. 1). The easiest way to reduce power is to insert a resistor in series with the lamp, adjusting its value to produce 50-watts dissipation in the lamp. If you go through the calculations, you will find that this *class-A* dimmer consumes about 20.7 watts, making its efficiency 70.7 percent.

To make a class-D dimmer, you would use a switching transistor instead of the resistor (fig. 1C). The transistor would be driven so that it is turned on 50% of the time and off the other 50%. The lamp will then consume an average of 50 watts and the transistor switch will consume (almost) nothing, thus having an efficiency of (almost) 100%. Note that the switch must

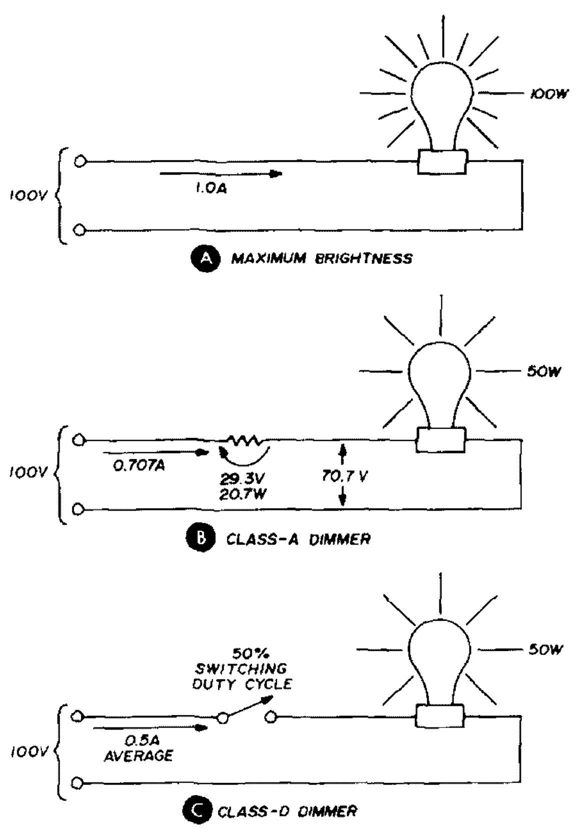


fig. 1. Simple class-A and class-D light dimmers illustrate relative efficiency of the two classes of operation. In the class-A light dimmer (B), maximum efficiency is 70.7%. Efficiency is nearly 100% in the class-D dimmer (C).

operate fast enough so that you cannot see the lamp flicker.

A technique which is useful when designing linear high-efficiency rf amplifiers is *envelope elimination* and *restoration*. The EER technique (fig. 2) was developed by Kahn in the early 1950s as a means of adapting high-power a-m transmitters to ssb service.[1,2] Basically, this technique treats a single-sideband signal as a carrier which is both amplitude and phase modulated. Before amplification, the envelope (amplitude) is detected and the carrier amplitude is limited. The phase-modulated, constant-amplitude carrier may now be amplified by non-linear rf amplifiers. The envelope signal is an audio signal with frequency components from dc to about 10 kHz (four times the tone separation will do), and can be amplified by linear audio amplifiers. The two signals are combined by amplitude modulating the last stage of the rf amplifiers, producing an amplified ssb signal. This technique is useful in high efficiency amplification since it is generally much easier to build linear class-D audio amplifiers than linear class-D rf amplifiers.

In this article I will attempt to give you a basic understanding of what class-D operation is all about. Understanding class-D requires close attention to where current flows and why a voltage is what it is. Never assume a particular waveshape just because it seems reasonable; make sure something causes it.

class-D audio amplifiers

Class-D audio amplification is often referred to as *pulse-width modulation* (PWM) because this technique is used in class-D audio amplifiers. Applications include both amplitude modulators and loudspeaker amplifiers, and all power levels of interest to amateurs are feasible. The use of integrated circuits makes the modulation circuitry quite simple.

Audio-frequency PWM is somewhat similar to the class-D light dimmer. A combination of transistors and diodes acts as a switch which applies a voltage pulse train to the load through a lowpass

filter. The pulse width determines the average voltage, which is the output of the lowpass filter. The output (average) voltage can be varied to reproduce a desired audio signal by varying the pulse width.

a-m modulator

When a class-D audio amplifier is used as an a-m collector modulator, both the voltage and current outputs will always be positive. The switch, composed of a transistor and a diode (**fig.** 3A), connects the inductor in the lowpass filter to either $+V_{cc}$ or ground (zero voltage). The lowpass filter has a high input impedance at the switching frequency, but allows the desired audio frequencies to reach the load. Load resistor R represents the rf amplifier to be modulated.

If the pulse width is 50% of the period of the switching frequency, the average voltage is 50% of V_{cc}. Since the lowpass filter allows only low-frequency voltages to reach the load, with a 50% pulse width a current of $V_{cc}/2R$ flows in the load. As the pulse width is *varied from zero* to 100%, the load voltage varies from zero to V_{cc}, and the load current varies from zero to V_{cc}/R.

When the transistor is driven *on*, current flows through the inductor into the load; when the transistor is turned *off* it stops conducting current. The inductor resists any instantaneous change in current flow and draws what it needs through the diode. Therefore, the diode acts as part of the switch and cannot be eliminated from the circuit. Note that no current at the switching frequency flows into the load; hence, there is no power dissipated at the switching frequency. When the transistor or diode conducts current, the voltage across it is very small (saturation voltage), so the switch is very efficient.[3,4,5]

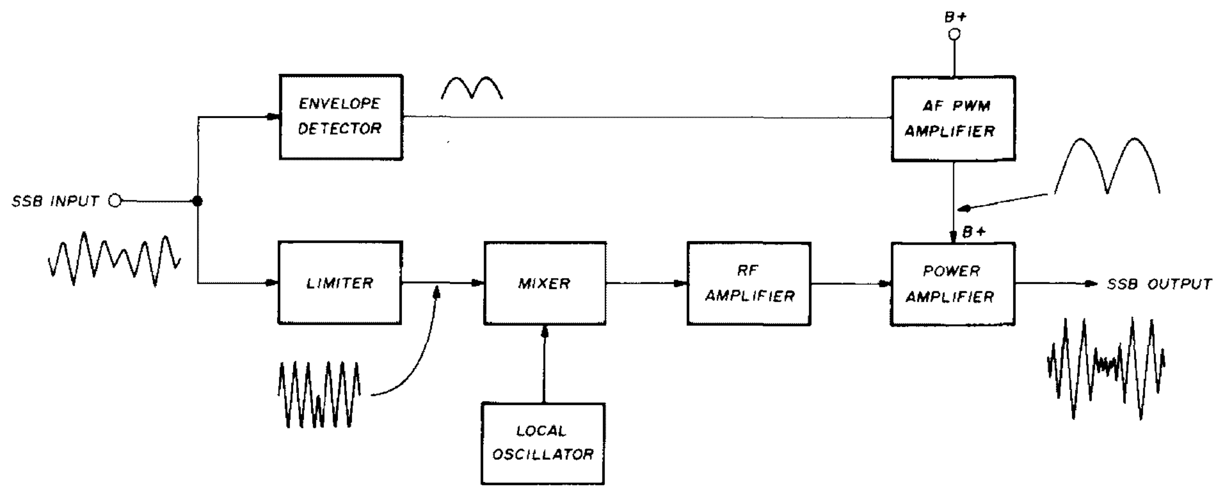


fig. 2. Block diagram of the basic envelope elimination and restoration system. This technique permits the use of non-linear rf power amplifiers in ssb service.

loudspeaker amplifier

If class-D is to be used in an audio power amplifier or other ac-coupled application, the output circuit becomes more complicated as shown in **fig. 4**. A blocking capacitor, C_B, prevents dc current from flowing into the load but requires that current be able to flow from the load into the switch during any part of the ac cycle. This requires a second transistor-diode pair. A pulse width of 50% produces no current flow at all. As the pulse width is varied, low-frequency current flows into the load.

A third variation of the output stage is sometimes called the class-BD amplifier[6] (the previous configurations are then called class-AD). The class-BD circuit is essentially both a positive and a negative voltage modulator (fig. 5). This configuration is very versatile because it can produce both ac and dc outputs. Also, when there is no signal input, the ampli-

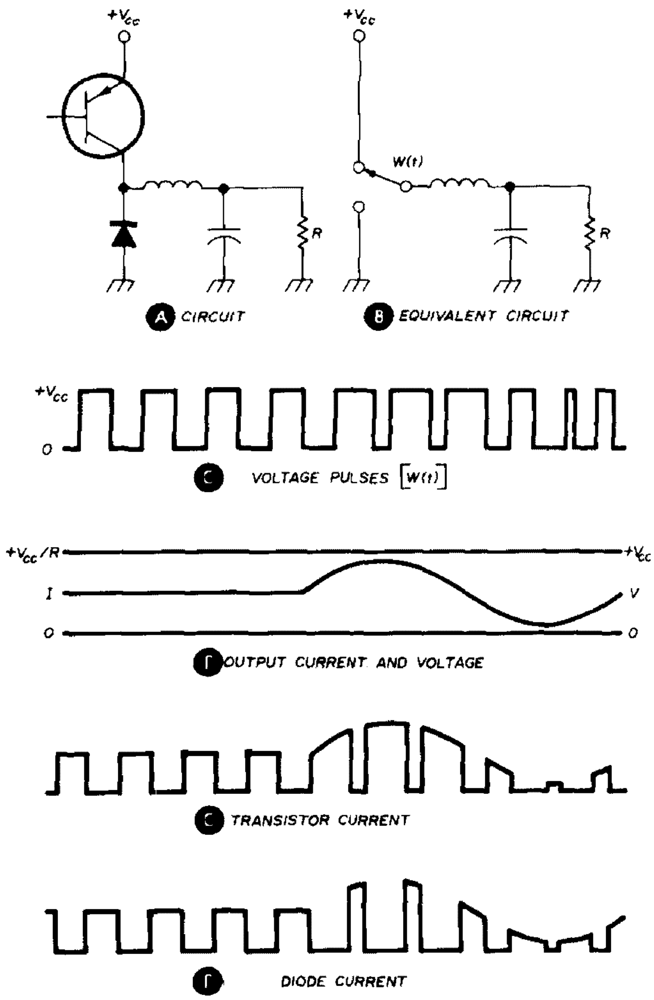


fig. 3. Basic audio-frequency pulse-width modulation circuit with voltage and current waveforms.

fier can simply shut down and conserve drive power. However, the driving and modulation circuitry is proportionately more complex, and two power supplies are required. Diodes CR1 and CR2 prevent conduction by the collector-base junctions of transistors Q1 and Q2 when Q3 and Q4 are turned *on*.

circuit parameters

In class-D audio power amplifiers the switching frequency must be somewhat higher than the highest audio frequency desired so the lowpass filter can eliminate switching frequency energy from the output. A ratio of 10:1 is nice, but a ratio of 5:1 (sometimes a little less) will do. In no case can the switching frequency be less than twice the highest audio frequency.

A pulse train contains a dc component, the switching frequency and its harmonics. As the pulse width is modulated, the dc component is modulated *linearly to produce the desired* output signal. The switching frequency and its harmonics are also modulated, but their modulation is highly nonlinear, resulting in a large amount of splatter near these frequencies. When the switching frequency is much higher than the audio frequency the splatter is reduced to negligible levels at the passband of the output filter.

The input impedance of the low-pass filter should be much higher than the load resistance to prevent significant current flow at the switching frequency. A good way to accomplish this is to use a large inductor at the input of the filter — its reactance, at the switching frequency, should be ten times the load resistance.

If the class-D amplifier is used as a loudspeaker amplifier, the output filter should attenuate the switching frequency by at least 40 dB. If the circuit is used as an a-m modulator, the filter should attenuate the switching frequency 80 dB to prevent adjacent-channel interference. A two-pole, lowpass filter will provide 40-dB attenuation per decade; a four-pole filter will provide 80-dB per decade attenuation.

It is convenient to make the cutoff frequency of the filter equal to the highest audio frequency and the switching frequency ten times higher; this allows 40-dB attenuation of the switching frequency for each two poles (L-C pair) in the output filter. The inductor value is chosen to have a reactance of 10R at the switching frequency and the capacitor is chosen to resonate with the inductor at the cutoff frequency. The capacitor can be any good quality capacitor. The inductor should use a toroid which has relatively high Q at the switching frequency.

generating PWM

Pulse-width modulation can be accomplished quite easily using only a couple of integrated circuits. **Fig. 6** shows how a triangle wave can be compared to the audio input signal to generate pulses *with*

widths proportional to the audio input. Whenever the input audio signal is greater than the triangle signal, the output pulse amplifier is driven *on*.

A good triangle wave of about 1-volt peak-to-peak can be generated by integrating a 12-volt peak-to-peak square wave with a lowpass R-C filter. The square wave can be generated by a simple multivibrator built from cmos NOR gates as shown in fig. 7. A high-speed comparator such as the 710 is then used to produce the pulses (fig. 8).

An alternative means of generating the pulses is to use a NE555 timer IC. The circuits are described in the applications notes.[7] This modulator varies only the trailing edge of the pulse, but that really doesn't matter much. For linear PWM, however, you must charge the modulator capacitor from a current source rather than through a resistor. A recent paper by Subbarao[8] describes a simple PWM circuit using a unijunction transistor to generate the triangle wave (if you try his circuit, I suggest using the diodes in the output as described here).

pulse amplifiers

Switching (pulse) amplifiers can be designed quite easily since there are only three basic considerations:[9]

1. There must be enough base current to saturate the transistors.

2. There must be enough drive voltage to cut off the transistors.

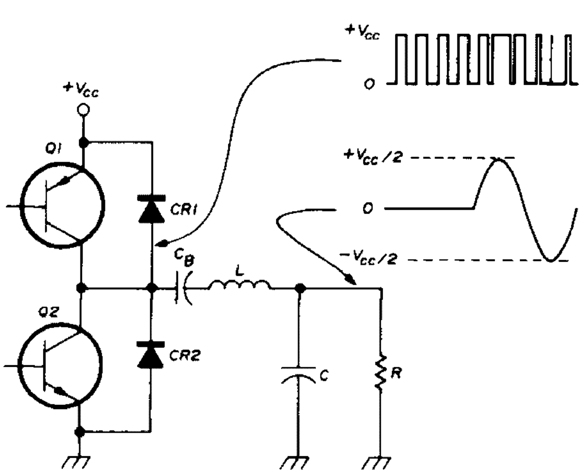


fig. 4. Basic class-D audio power amplifier (also called class AD).

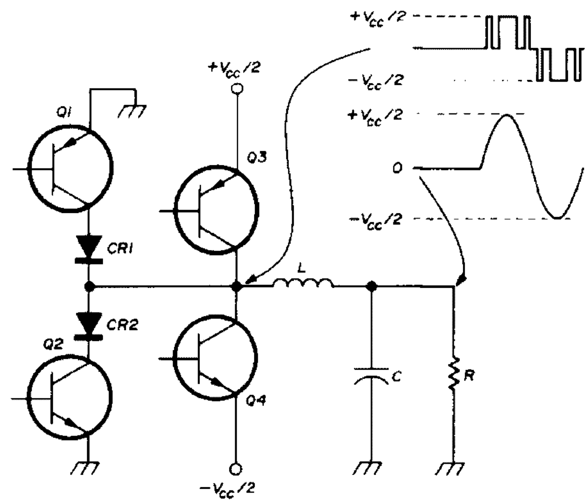


fig. 5. This class-D amplifier (also called class BD) can produce both ac and dc outputs.

3. Rise time must be much smaller than the switching frequency.

The third requirement is satisfied by using transistors with a gain-bandwidth product at least ten times the switching frequency (or transistors with a rise time no larger than 5% of the switching period). There are many inexpensive switching transistors suitable for low and medium power levels.

Pulse amplifiers are best designed by working from the output stage back to the pulse source. A sample design is shown in fig. 9. Assume that the peak output current is 1 ampere, and that the minimum current gain of transistor Q3 is 25 (from the data sheet). The base current required to saturate Q3 at all times is then

$$i_{b3} = \frac{i_{c3}}{\beta} = \frac{1A}{25} = 40 \text{ mA}$$

The output transistor is to be driven by the complimentary transistor pair, Q1 and Q2. Under saturated conditions, silicon transistors have typical collector-emitter voltages of 0.3 volt and typical base-emitter voltages of 0.7 volt. Transistor Q2 must be *on* to turn Q3 *on*, so the voltage at point D will be 0.3 volt. The voltage on the base of Q3 (point E) will be 0.7 volt below V_{cc}, or 11.3 volts. The maximum value of R3 is then

$$R3_{max} = \frac{11.3 - 0.3}{0.040} = 275 \text{ ohms}$$

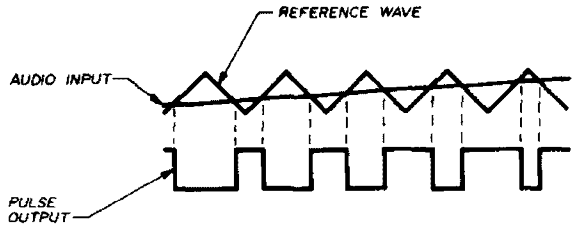


fig. 6. A PWM signal can be generated by comparing a sinusoidal input signal to a reference triangle waveform. A suitable triangle waveform generator is shown in fig. 7; a PWM generator using an IC comparator is illustrated in fig. 8.

Use the nearest standard value (270 ohms). When transistor Q1 is *on* and Q2 is *off*, point D will be at 11.7 volts (12-0.3). Since this is greater than the 11.3-volt saturation voltage, transistor Q3 should be cut off.

A similar process is used to choose a value for R2 (6800 ohms), assuming that point A is also driven by a complimentary pair. Note that although transistor Q1 does not conduct any sustained current, it will conduct the same surge current while switching, so R1 should be the same value as R2. A complimentary pair (such as Q1 and Q2) will generally be necessary for high switching speeds. If you use a single transistor and collector resistor, turn-on will be fast. However, turn-off will be slow because the collector capacitance must discharge through the resistor.

An emitter follower pair cannot provide any voltage gain, and will produce a smaller voltage swing in the output, and thus be less efficient. For these reasons a voltage gain pair is usually better. However, there are disadvantages. If point A is disconnected from its drive, both Q1 and Q2 will turn on, causing a short circuit, probably destroying themselves and Q3 as well. During switching this same process can cause a large current spike, reducing efficiency. It is generally a good idea to put a small resistor (1% to 5% of R3) somewhere in the collector path to reduce these effects.

Transistor base capacitance can be a significant impediment to high-speed switching. Before a transistor can turn on, its base capacitance must be charged to 0.7 volt through the base resistor. This produces an output waveform delayed from the driving waveform. This effect can be overcome by using speed-up capacitors (C1, C2 and C3 in **fig. 9**). These capacitors have much the same effect as the trimmer capacitor in an oscilloscope probe. The correct capacitance value causes a square-wave voltage (rather than a damped voltage) to appear on the base.

It is probably possible to calculate the values of these capacitors, but it is much easier to find the values by trial and error. Use an oscilloscope to compare the rising edges of the square waves at points A and D. Install capacitors C1 and C2 and observe that the delay becomes smaller; increase the value of C1 and C2 until the delay is negligible. Do not use more capacitance than is necessary as this will load the driver and slow the amplifier. Base-to-emitter resistors (such as R5) may be added to provide return paths for the speed-up capacitors. This helps to keep the transistors *off* when not driven *on*. Typical values (determined experimentally) are one-half to one-third of the corresponding base resistors.

One last comment on pulse amplifiers. Fixed voltage changes can sometimes be made by using diode drops. The diode(s) will conduct only when the applied voltage is large enough.

efficiency

What efficiency can you actually expect? Both saturation voltage and non-instantaneous switching reduce the efficiency of the output stage, and these and current spikes reduce the efficiency

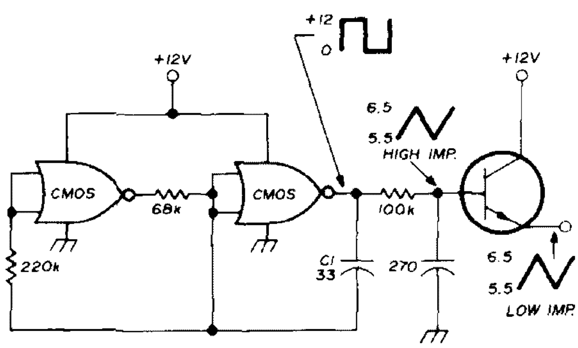


fig. 7. Simple reference triangle wave generator uses two cmos gates and one transistor.

of the driver stages. The driving resistors also consume power. Although exact calculation of efficiency requires some complicated formulas, here are some rules of thumb that give ballpark answers:

1. Decrease the efficiency of the output stage by subtracting the saturation voltage-to-V_{cc} ratio from 1. If the satura-

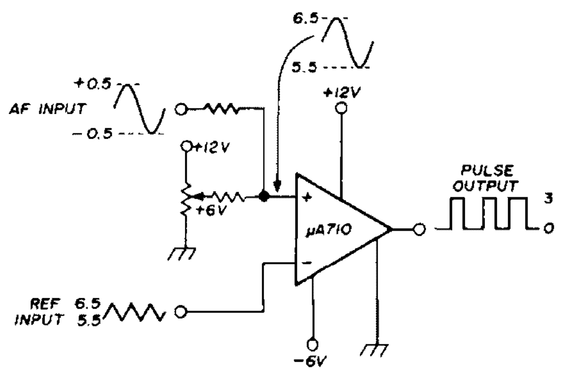


fig. 8. Simple IC comparator circuit generates PWM signal. Reference triangle waveform is provided by circuit of fig. 7. Input and output waveforms are shown in fig. 6.

tion voltage is 0.3 volt and the supply is 12 volts,

$$\eta = 1.00 - \frac{0.3}{12} = 1.0 - 0.025 = 97.5\%$$

2. Further decrease the efficiency by subtracting the ratio of the time spent in switching to the switching period. If the switching frequency is 100 kHz (10 μs) and the rise time is 100 nanoseconds,

$$\eta = 0.975 - \frac{2 \cdot 100 \text{ nsec}}{10\ \mu\text{sec}} = 95.5\%$$

3. Most of the power dissipated in the amplifier will be dissipated in the output stage and its base driving resistor(s). The power consumed in the base resistor is approximately equal to the output power reduced by the ratio of base current to output current. This decreases the efficiency by the ratio of i_b to i_c (or by $1/\beta$)

$$\eta = 0.955 - \frac{41 \text{ mA}}{1 \text{ A}} = 91.4\%$$

Remember that these rules are only approximate, so the actual efficiency will be lower at less than peak output. Before you get discouraged, though, remember that saturation voltage and drive power reduce class-B efficiency, and that peak power contributes most to the average efficiency.

power regulators

Class-D operation can also be used for ac and dc power control. SCRs and triacs can be used to chop an ac waveform to produce the desired output power. These devices turn on when driven and stay on until the output current ceases (at the end of an ac half cycle). SCRs conduct current in only one direction, while triacs conduct current in either direction.

Dc voltages (or currents) can be regulated efficiently by switching regulators.[10,11] A voltage-reducing regulator circuit looks like a simplified audio amplifier (fig. 10). The high-gain amplifier compares the output of the lowpass filter to the desired reference voltage. If the output voltage is lower than the reference by more than a small amount, transistor Q1 is turned on and the output voltage begins to rise. When the output voltage exceeds the reference by more than a small amount, Q1 is switched off and the output begins to fall. In this manner the output voltage is kept within a small fixed amount of the reference voltage. The switching frequency depends on the output filter and current, and varies with changing load conditions. The high-gain amplifier and voltage reference

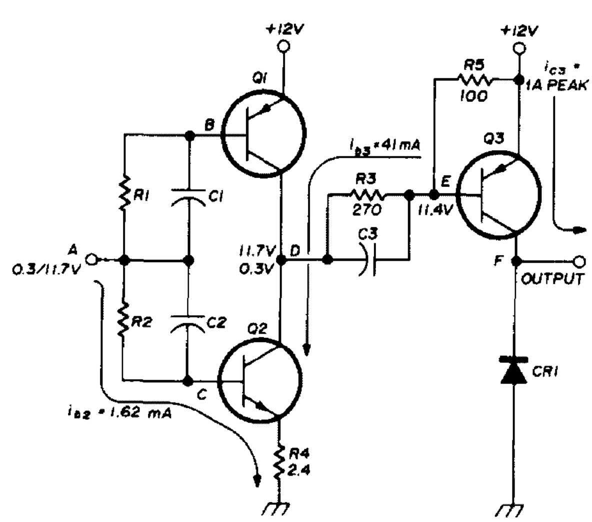


fig. 9. Basic pulse amplifier circuit. Design and operation of this circuit is discussed in the text.

are often available in a single IC such as the μA723.

class-D rf amplifiers

Class-D operation can also be used to build high-efficiency rf amplifiers.[12] In this case the switching frequency component of the pulse train becomes the rf carrier and is coupled to the load through a bandpass filter rather than a lowpass filter. **Fig. 11** shows the equivalent circuit of a simple class-D rf amplifier. Switches S1 and S2 (switching transistors in an actual circuit) are driven alternately on and off with a 50% duty cycle, producing a square-wave voltage, V_s. The square-wave voltage has components at the fundamental (switching) frequency and its odd harmonics. The peak-to-peak voltage of the fundamental component is $4/\pi$ (1.27) times the peak-to-peak voltage of the square wave. The series-tuned output circuit is resonant at the switching frequency, so it exhibits high impedance to the harmonics, but provides a direct connection to the load for the carrier. As a result, only current at the carrier frequency can flow into the load.

The output voltage and current are sine waves which are represented mathematically by

$$v_o = \frac{2V_{cc}}{\pi} \sin \omega t$$

$$i_o = \frac{v_o}{R} = \frac{2V_{cc}}{\pi R} \sin \omega t$$

When the output current, i_o, is positive, it flows through S1; when it is negative, it flows through S2. A single transistor switch cannot be used for class-D rf operation because the output current must be able to flow in both directions. The power output from this amplifier is given by

$$P_o = \frac{(v_{o\ peak})(i_{o\ peak})}{2} = \frac{2V^2_{cc}}{\pi^2 R}$$

The switches consume almost no power themselves because the voltage across them is nearly zero when they are conducting current.

rf circuits

There are a variety of class-D rf amplifier circuits (**fig. 12**) just as there are a wide variety of linear amplifiers. The equivalent circuit just discussed can be realized by any of the complimentary versions (**figs. 12A, 12B** or **12C**). The input transformer secondary windings are

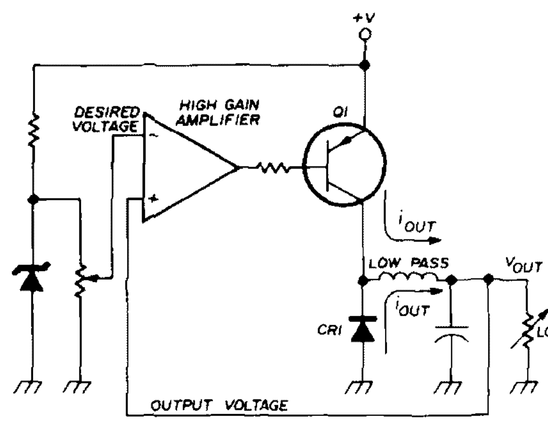


fig. 10. Basic switching voltage-regulator circuit.

connected so that the transistors turn on with opposite polarities of the drive signal. Complimentary class-D amplifiers have the advantage of requiring no output transformer, and hence no transformer losses.

The quasi-complimentary configuration (**fig. 12C**) uses only npn transistors which are both more abundant and less expensive than pnp types at high frequencies. This is a definite advantage of the quasi-complimentary circuit.

Transformer-coupled or push-pull class-D amplifiers are also possible. In the voltage-switching amplifier (**fig. 12D**), alternately driving the transistors *on* causes their collector voltages to be square waves of zero to $2V_{cc}$ volts. The *current* which flows in either transistor is a half cycle of a sine wave, just as in the complimentary amplifiers. Power output depends on the turns ratio of the output transformer, as well as on the supply voltage and load resistance.

The total voltage swing on the transformer primary is $2V_{cc}$, so if there are four turns in the primary for each turn in the secondary, the output power will be

the same as that of a complimentary amplifier. Output voltage varies with the turns ratio, and output power varies with the square of the turns ratio. The turns ratio can be used as part of the matching and loading network. This configuration is particularly useful when the supply voltage is low since it provides a $2V_{cc}$ voltage swing at the collectors of the power transistors.

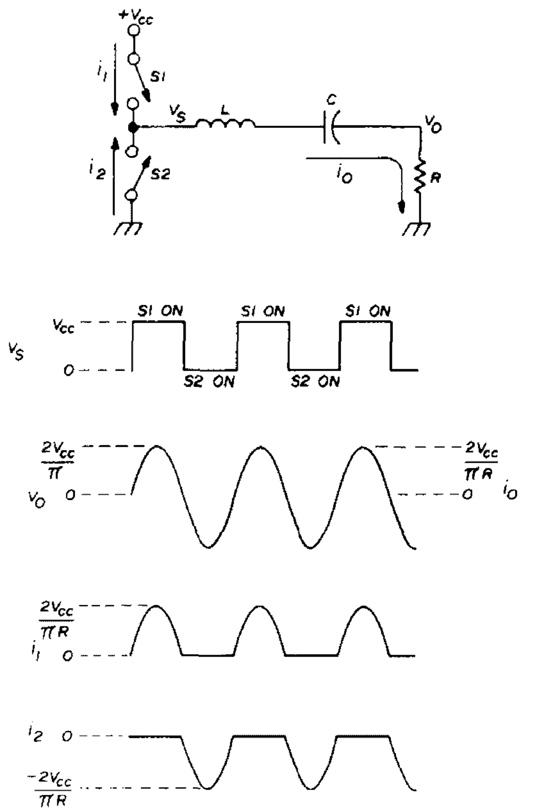


fig. 11. Equivalent circuit of a class-D rf amplifier and ideal waveforms.

This is a good time to point out that you cannot connect an rf choke directly to the collector or primary center tap in a voltage-switching amplifier (**fig. 12A, 12B, 12C** or **12D**). These amplifiers draw current in lumps at the signal frequency, and an rf choke tries to force constant current. A capacitor should be connected from either the collector of transistor Q1 or the primary center tap to ground to supply current as needed while maintaining constant voltage. An rf choke can then be connected from the collector to the supply to keep rf out of the power supply.

Another class-D rf circuit is the current-switching push-pull amplifier shown in **fig. 12E**. This circuit is a so-called "dual" of the voltage-switching amplifier because it features square-wave current and sine-wave voltages. The rf choke forces a constant current so the transistors conduct constant current while they are *on*. The parallel-tuned output filter conducts all of the harmonic currents but forces the carrier current into the load. Output voltage and power are somewhat different than those of the voltage-switching amplifier. Since there can be no dc voltage drop across a transformer winding, the total primary voltage must be

$$v_{pri} = \pi V_{cc} \sin \omega t$$

(The peak collector voltage is πV_{cc} so the average or dc collector voltage is V_{cc}.) If the turns ratio is 1:1, the output power is

$$P_o = \frac{\pi^2 V^2{}_{cc}}{2R}$$

Output current is

$$i_o = \frac{\pi V_{cc}}{R} \sin \omega t$$

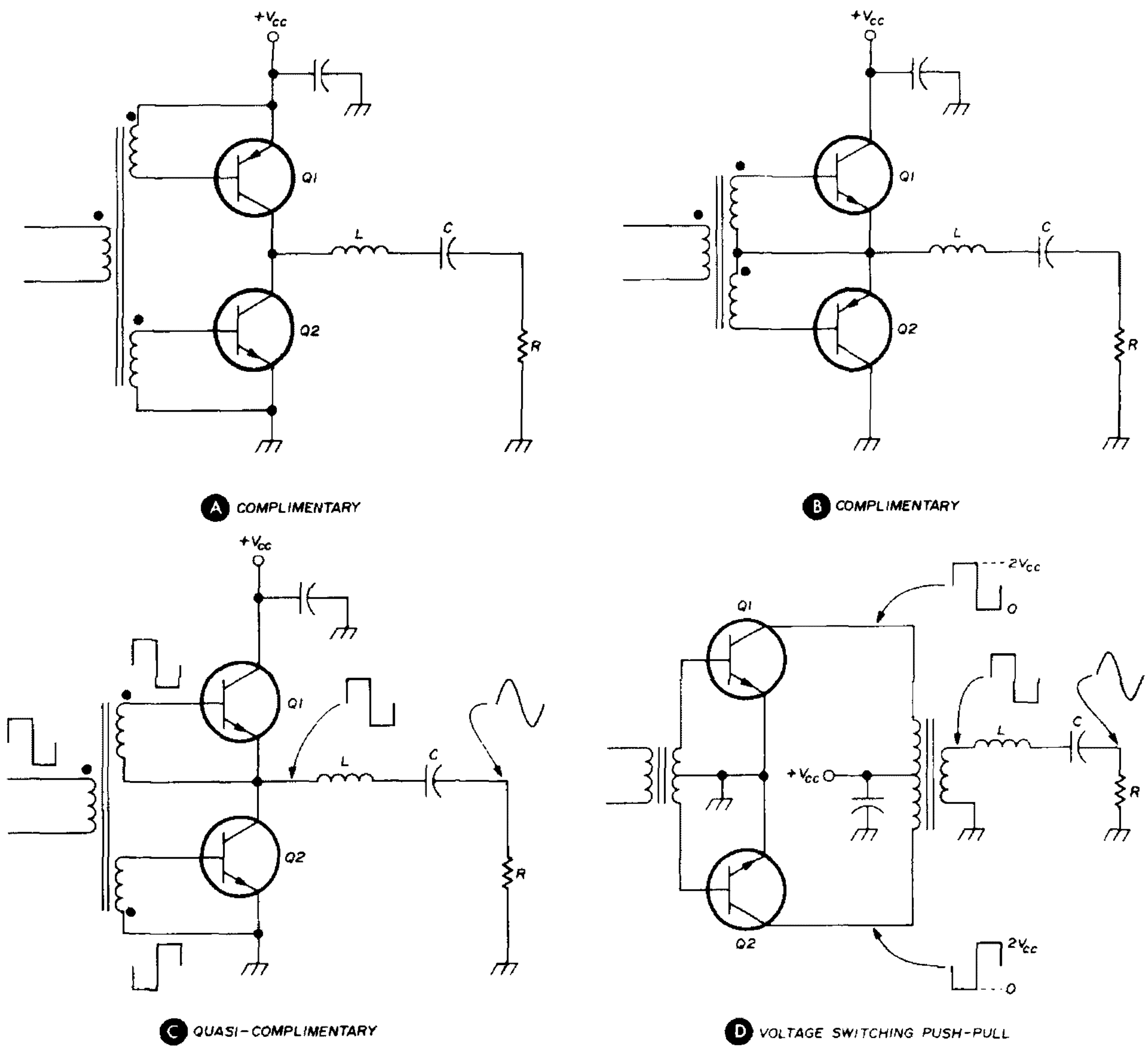
and the dc input current (and transistor peak current) is

$$I = \frac{\pi^2 V_{cc}}{2R}$$

The current-switching amplifier offers an alternative way of fitting together power requirements, supply voltage and maximum transistor ratings. Also, it has the same configuration and the same output filters as class-B transistor power amplifiers. However, it will generally be less efficient than the voltage-switching amplifier because more current must be drawn through the same saturation voltage and large amounts of current must be switched (the voltage-switching amplifier switches at zero current points).

transformers and tuned circuits

The transformers required in class-D rf amplifiers are essentially the same as



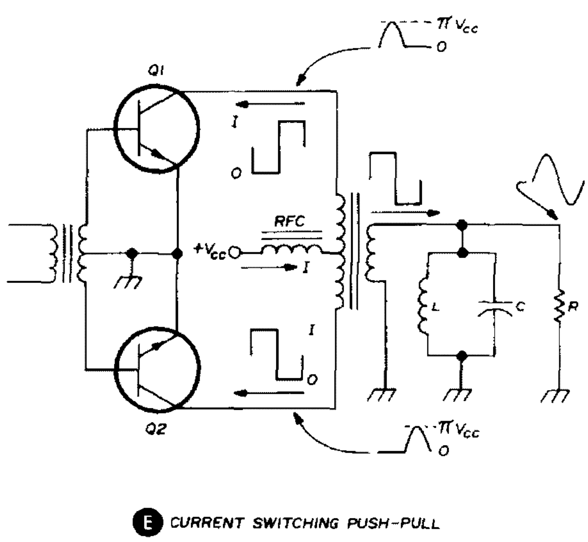


fig. 12. Practical class-D rf amplifiers can be built in various ways as shown here. The advantages and disadvantages of each of these circuits are discussed in the text.

those used in solid-state class-B rf amplifiers.[15] They consist of several parallel wires wound on a stack of toroidal cores made of suitable material for the frequencies to be used. Building these transformers is a mixture of intuition and witchcraft and a lot of trial and error. However, here are some general rules:

1. Increasing the number of turns increases the series inductance, degrading the high-frequency response.

2. Paralleling two windings with the same number of turns decreases series inductance, improving high-frequency response.

3. Stacking several cores increases power handling capability and increases shunt inductance, improving low-frequency response.

The transformer parameters must be chosen to fit the application, and in class-D circuits it is necessary to pass a few harmonics of the carrier as well as the carrier itself. However, it is not difficult

to build transformers working over several octaves (up to 5 or 6).[16,17]

The output filter need not be the simple tuned circuits used in the examples here. A T-network presents a high impedance to the harmonics of the carrier, and can be used where series-tuned circuits are shown. Pi-networks present low impedances to the harmonics and can be used in the place of parallel-tuned circuits. Input impedance to the third harmonic should be at least ten times the load impedance at the carrier frequency, and the output filter should attenuate harmonics by 50 dB or more.

Two (or more) small amplifiers can be combined to obtain greater output power by connecting them to the load through a hybrid transformer (fig. 13) as is done with linear amplifiers.[18] Basically, a hybrid transformer prevents the two amplifiers from loading each other. The transformer has a 1:1 turns ratio which forces equal currents in each winding. If you wish, you can work out the details with a little algebra, but essentially, if the two amplifiers do not produce equal outputs, R2 consumes power and reduces efficiency. This means the two amplifiers must be closely matched or controlled.

Hybrid transformers are also used to divide drive power between two amplifiers whose outputs will be combined. This helps to keep phase relationships the same in each amplifier. If more than two amplifiers are to be combined, they *must*

table 1. Transistor types that are suggested for use in high-efficiency class-D amplifiers. However, these are not the only ones that can be used. Avoid using rf transistors which are optimized for particular frequencies as they usually contain internal matching which usually won't help them switch fast. These specs are intended as a general guide rather than exact specifications.

JEDEC number	type	maximum V_{CE}*	maximum I_C	package	application
2N4123	npn	30	0.2	TO-92	100-kHz switch
2N4124	npn	25	0.2	(plastic)	
2N4125	pnp	30	0.2		
2N4126	pnp	25	0.2		
2N2222A	npn	40/65**	0.8	TO-18	1-MHz switch
2N2907A	pnp	40/65**	0.8	(metal)	
2N5190	npn	40	4.0	flat	100-kHz switch
2N5191	npn	60	4.0	(plastic)	
2N5192	npn	80	4.0		
2N5193	pnp	40	4.0		
2N5194	pnp	60	4.0		
2N5195	pnp	80	4.0		
2N3262	npn	80/100	1.5	TO-39	
2N3734	npn	30/45	1.5	TO-5 (metal)	30-MHz switch (high-speed switch)
2N3735	npn	50/70	1.5		
2N5262	npn	50/70	3.0	TO-39	
2N3961	npn	40/56	1.0	stud	30-MHz switch (400-MHz rf)
2N3553	npn	40/56	1.5	can	
2N3375	npn	40/56	1.5	stud	
2N3632	npn	40/56	3.0	stud	
SD-200	fet	25	0.05	can	100-MHz switch
1N4933	diode	50	1.0	DO-41	100-kHz switch
MR380	diode	50	3.0	(metal)	

*When two values are shown they are V_{CBO}/V_{CEV}. The first is with base open, the second with reverse bias as in a push-pull amplifier.

**Without A suffix, 30/50.

be combined in groups of two. The drive signal is first split into two signals, then into four. Two pairs of outputs are combined to form two signals; these signals are then combined to form one high-power output. Note that the input impedance of the hybrid transformer and the difference load resistor are twice the output resistance.

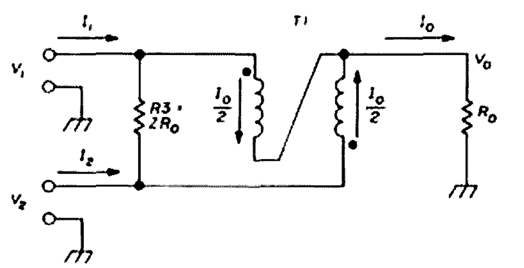


fig. 13. Hybrid transformers, such as the one shown here, can be used to combine the outputs from two amplifiers.

class-D rf amplifier design

Designing a class-D rf amplifier is generally easier than designing a class-B or -C rf amplifier stage. There are no gain-variation, base-bias, neutralization or unequal-current problems to worry about. As with class-D audio amplifiers, you start at the output and work backwards.

I decided to try to design and build a 50-watt class-D power amplifier for 160 and 80 meters (and maybe 40). The circuit that resulted is shown in fig. 14. The first step is selecting the Q of the output circuit. I chose a Q of 5 which meant that L3 and C6 must have reactances of 250 ohms at the midband frequency. This value of Q reduces the third harmonic to 33 dB below the carrier.

The next step is to determine the output voltage and current. The peak voltage and peak current into the 50-ohm load are found from

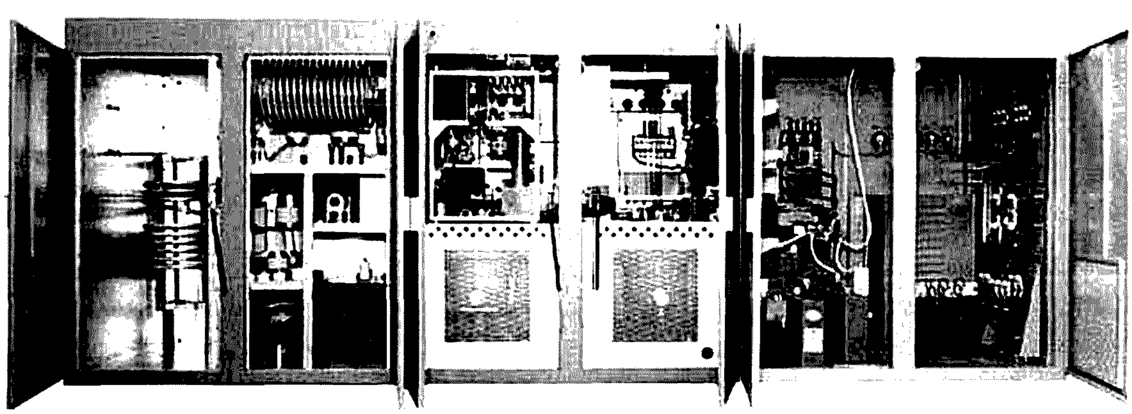
$$v_p = \sqrt{2RP} = \sqrt{2 \cdot 50 \cdot 50} = 70.7 \text{ volts}$$

$$i_p = v_p/R = 70.7/50 = 1.41 \text{ amp}$$

The peak value of square-wave voltage at the input of the tuned circuit which will produce this peak output is

$$v_{swp} = \frac{\pi}{4} \cdot 70.7 = 55.5 \text{ volts}$$

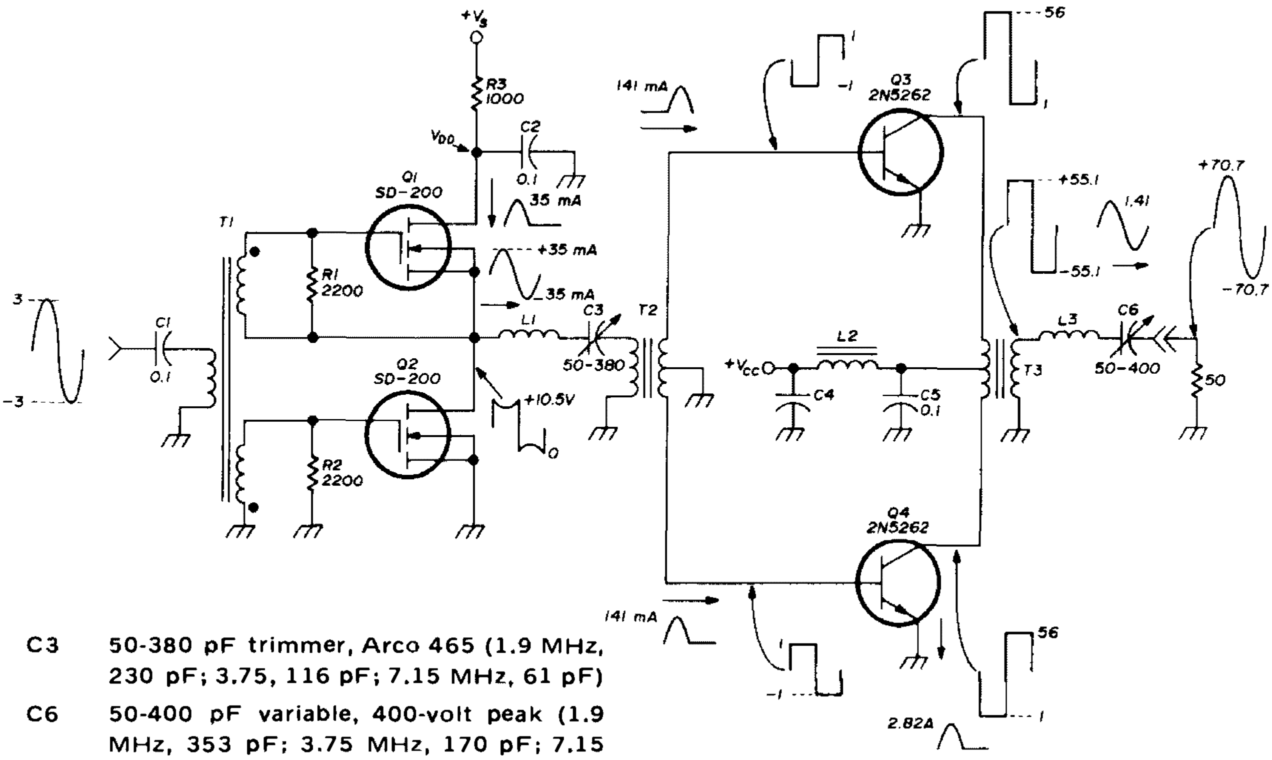
If no matching were used, the transistors would have to withstand 111 volts or



The Gates Radio Company VP-100 is a 100-kW a-m broadcast transmitter which uses a class-C rf amplifier and a class-D modulator. The modulator switches at 75-kHz and has an efficiency of about 90%. Overall efficiency is about 65%. The rf output circuitry is on the left, the PA and modulator are in the middle, and the power supplies are on the right.

more on their collectors. It is convenient to use a 4:1 matching transformer which reduces the square-wave voltage swing to 27.8 volts and increases peak current to 2.83 amps. These values are much more suitable for commercially available transistors.

I decided on the push-pull circuit rather than the quasi-complimentary configuration because I was already using a matching transformer and because a 28-volt power supply is easier to come by than a 56-volt supply. Capacitor C5 provides an ac ground for the transformer center tap,



C3	50-380 pF trimmer, Arco 465 (1.9 MHz, 230 pF; 3.75, 116 pF; 7.15 MHz, 61 pF)	
C6	50-400 pF variable, 400-volt peak (1.9 MHz, 353 pF; 3.75 MHz, 170 pF; 7.15 MHz, 89 pF)	
L1	160 meters	63 turns no. 30 on Permacore 57-1753 core (30.5 µH)
	80 meters	44 turns no. 26 on Permacore 57-1753 core (15.4 µH)
	40 meters	36 turns no. 26 on Permacore 57-1677 core (8.1 µH)
L2	16 turns no. 26 on Permacore 56-3596 ferrite core (1.0 µH)	
L3	160 meters	52 turns no. 26 on Permacore 57-1753 core (20.9 µH)
	80 meters	42 turns no. 26 on Permacore 57-1677 core (10.6 µH)
	40 meters	30 turns no. 26 on Permacore 57-1677 core (5.6 µH)

T1	primary is 5 turns no. 26 enamelled on Ferroxcube 226T125-3E2A ferrite core. Secondary windings are each 25 turns no. 26 on same core
T2	primary is 8 turns no. 20 enamelled wire wound through 6 Ceramic Magnetics CN-20 cores (two parallel stacks). Secondary is 4 turns no. 20, center-tapped, through same cores (see text)
T3	primary is 4 turns no. 20 enamelled wire, center-tapped to C5, wound through 12 Ceramic Magnetics CN-20 cores (two parallel stacks). Secondary is 4 turns no. 20 through same cores (see text and fig. 15)

fig. 14. Class-D rf power amplifier for 160, 80 and 40 meters provides more than 35 watts output with collector efficiencies of 90% or more. Construction of transformer T3 is shown in fig. 15.

If the saturation voltage is 1 volt, the collector supply voltage, V_{cc}, must be 28.5 volts, and the collector voltage will swing between 1 and 56 volts. The dc current required is

$$i_{dc} = \frac{2i_p}{\pi} = \frac{2 \cdot 2.83}{\pi} = 1.80 \text{ A}$$

and the L2-C4 combination keeps rf out of the power-supply wiring.

Now the driving circuitry must be designed. It is convenient to drive the bases of transistors Q3 and Q4 through a push-pull transformer to insure that they are driven out-of-phase with respect to

each other. It is best to drive the bases with sine-wave current.

While square-wave current might turn the transistors *on* a little more rapidly, it maintains a lot of charge in the base which must be removed before the transistors will turn *off*. A check of switching transistor characteristics will show that storage time (turn-off delay) is a much more serious problem than turn-on delay. Sine-wave current provides the base with only enough charge to sustain the collector current, greatly reducing storage-time effects.

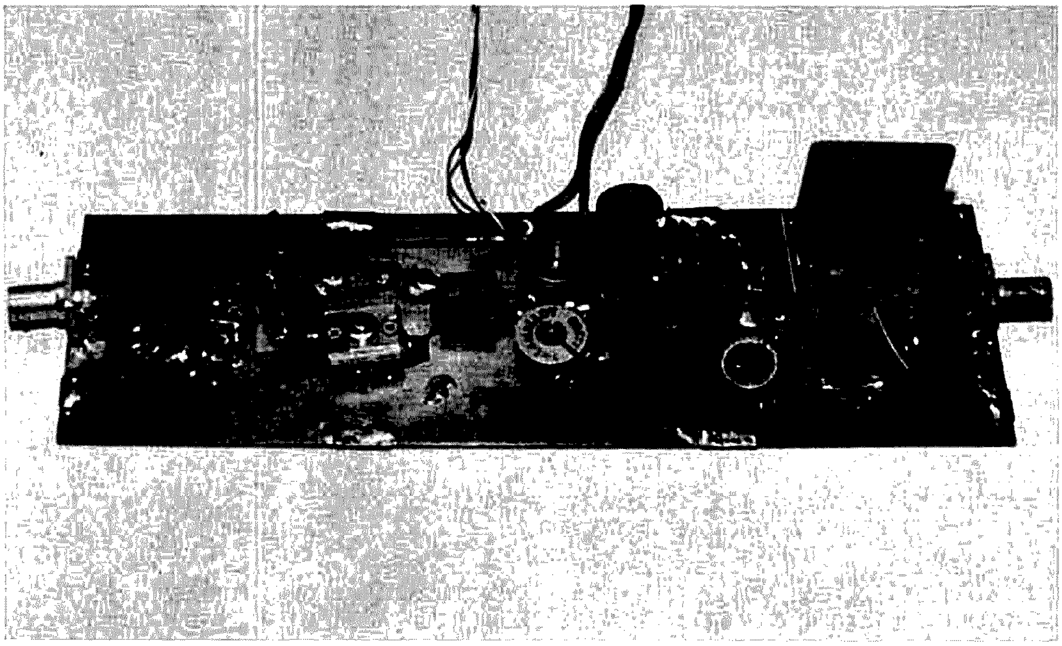
If the output transistors have a current gain of 20, to insure saturation the base current in each transistor must be a half sine-wave with a peak value of

$$i_{bp} = \frac{2.83}{20} = 142 \text{ mA}$$

The base voltage will rise to about 1 volt (or slightly less) when driven. When the current changes direction, the base becomes an open circuit and its voltage becomes −1 volt. This is a reflection of the +1 volt potential on the base of the other transistor through transformer T2. The power which must be applied to the two transistor bases is then

$$P_b = \frac{\frac{4}{\pi} \cdot 1 \cdot 0.142}{2} = 90 \text{ mW}$$

This means that the power amplifier has a power gain of about 27 dB!



Author's experimental class-D power amplifier. Parts are in the same general locations as in the circuit diagram (fig. 14). Note that C6 is elevated by pieces of PC board.

Since the drive power is only a small fraction of the total power this amplifier will require, driver efficiency is not too important. Almost any type of rf amplifier can be used for a driver — even untuned class-A. I found it more convenient to use a quasi-complimentary class-D amplifier. I used a pair of relatively new fets for Q1 and Q2. These have greater stability and faster switching speed than most bipolar devices, and are easier to drive as well.

The series-tuned circuit L1-C3 will allow only sinuosidal current to flow. Since harmonic suppression is not as important as in the power-amplifier stage, I used a Q of 2.5 for the driver. At the carrier frequency transistors Q3 and Q4 have a base resistance of

$$R_b = \frac{(4/\pi) \cdot 1}{0.142} = 8.97 \text{ ohms}$$

It would be difficult to build a 90-mW

driver with a 9-ohm load line, so T2 has a turns ratio of 4:1. This multiplies the impedance by 16, resulting in a 145-ohm load line. The ±1 volt base voltage will become ±4 volts on the input of T2. The current input to T2 has peaks of about 35 mA.

When saturated, the fets become 45-ohm resistors. This causes a voltage drop. To correct for this voltage drop, it is necessary to increase the supply voltage, V_{DD}. If there were no voltage drop through the fets, V_{DD} would be 8 volts to allow a ±4-volt swing. With the correction,

$$V_{DD} = 8 \cdot \frac{145 + 45}{145} = 10.5 \text{ volts}$$

The driver dc input current is 11.1 mA ($35/\pi$) so the driver power input is

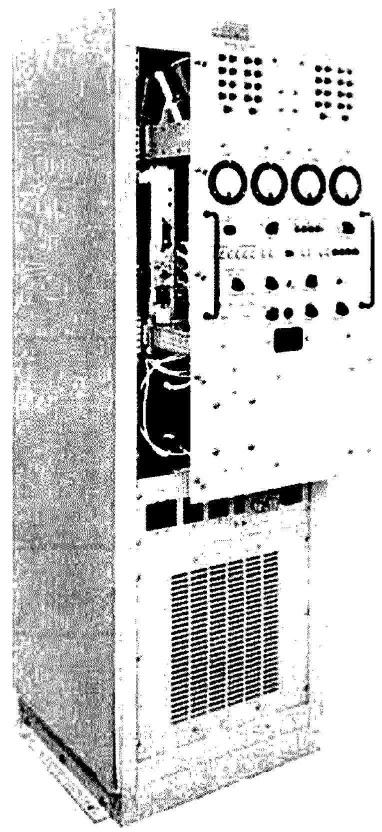
$$P_D = (10.5)\ (0.011) = 117 \text{ mW}$$

The driver efficiency is about 77%.

It is not a good idea to try to operate the driver from a 10.5-volt power supply. Remember that the value of 10.5 volts is related directly to the nominal 1-volt base saturation voltage. This was not an exact value to start with, and even if it were, it would vary slightly with temperature, collector current, etc. If V_{DD} were fixed at 10.5 volts and the base voltage decreased slightly, only the fet saturation resistance would limit the current. For this reason, it is better to use a higher voltage supply, V_s, and to provide a resistor to absorb changes in the base voltage. I used a 1k resistor and a 21.6-volt supply, which made it easy to monitor drive current. With this resistor the overall drive efficiency is about 37%, but the 240-mW total drive power is still less than one-percent of the power-amplifier power.

The fets are saturated by applying a 15-volt peak sinusoidal gate-to-source voltage. The 1:5 turns ratio of T1 permitted driving them from a 3-volt source. Since the fets have a very high input impedance and require essentially no input power, it was necessary to load the secondary windings to produce an approximate 50-ohm input impedance.

Transformers T2 and T3 are made by running enamelled wire through two parallel stacks of ferrite cores as shown in fig. 15. Transformer T3 has four turns in the output winding and two turns in each side of the input; T2 has two turns in each side of the output and eight turns in the input winding. Since I had the smaller size cores, twelve cores were required for



The Continental Electronics 314(E) is a 2-kW CW/MCW all solid-state transmitter for 275 to 530 kHz. It uses a push-pull class-D rf amplifier, and has an overall efficiency of 70% or better with load vswr as high as 3:1. Photo courtesy J.D. Rogers.

T3 and six for T2. The length of the individual cores is not important but the length of the transformer is, so if you have longer cores, use fewer of them. Both T2 and T3 show a phase angle of 10° or less on 160 and 80 meters. On 40 meters and up, series inductance appears. The series-tuned circuits can tune out this inductance on a particular frequency.

Two possibilities for Q3 and Q4 are the 2N5262, a switching transistor, and

the 2N3632, an rf transitor. Either transistor should handle the required current and voltage. Since 2N5262s were considerably cheaper, I decided to use them. It's my guess that rf transistors would be less efficient but more rugged. Clip-on heatsinks will reduce thermal stress on Q3 and Q4.

My amplifier was built on a double-sided PC board to insure a good ground plane. Layout is not critical, but remember that a 1.8-MHz class-D amplifier contains energy at frequencies much higher than 1.8 MHz.

The driver and power amplifier tuning coils are wound on *0.680-inch (17-mm)* OD rf toroid cores. All have a measured Q of 190 or more. The Q of these inductors is very important since it determines the amount of power they absorb from the output. The fraction of power lost to the coil is the ratio of circuit Q (5 for the output) to the inductor's unloaded Q. An inductor with a Q of 200, for example, causes a 2.5% reduction in efficiency. Remember that winding coils, like winding transformers, is a black art. Don't think you can improve the Q by using larger wire – you will decrease the self-resonant frequency and the Q because the interwinding capacitance is increased.

Test the driver first, with power removed from the power amplifier stage. Tune the driver for peak input current (if tuning for peak current seems strange, remember that this is series tuned, whereas most class-C power stages are parallel tuned). Input current should not exceed 16.6 mA, as this corresponds to the 50-mA peak current rating of the fet. Also, V_S should be kept at 25 volts or less. If the bases of Q3 and Q4 suddenly open, V_{DD} will go up to V_s. The driver should work on 160, 80 and 40 meters. It will also work on 20 if you reduce L1 to compensate for the series inductance of T2. You can tell when the input signal is large enough by whether any further increases produce increases in the dc current.

Now for the final. *Always* turn the driver on first (and off last). This circuit is not designed to have both Q3 and Q4 on at the same time, so it will oscillate, probably destroying the transistors. Start V_{cc} at zero and bring it up to 3 or 4 volts so you can measure the output. In the absence of an output meter, tune for peak dc current. Tune the power amplifier only at very low voltages. When the

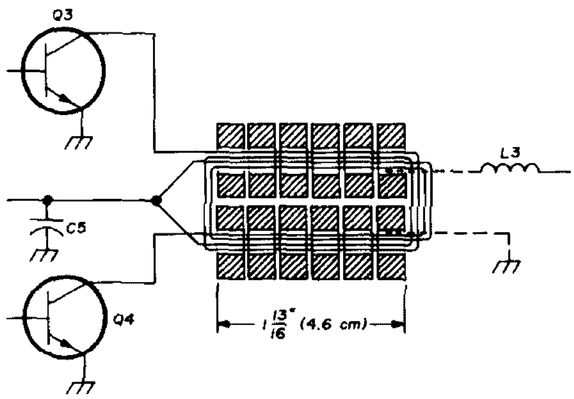


fig. 15. Layout of the output transformer (T3 in fig. 14). Primary is 4 turns no. 20 enamelled, center-tapped to C5, wound through 12 CN-20 ferrite cores (two parallel stacks). Secondary is 4 turns no. 20.

load is reactive (mistuned), large voltage spikes are generated, and these will usually destroy the transistors. After the stage is tuned up, increase V_{cc} gradually, Check to see that Q3 and Q4 are saturated by slightly varying the drive (V_S). If there is any variation in the output power or dc input to the final, it needs more drive.

I first tested the amplifier at the 10-watt level on all three bands. I measured collector efficiencies of 95%, 92% and 89% on 160, 80 and 40 meters, respectively. As you increase V_{cc}, efficiency will improve because the saturation voltage becomes less significant. The power supply I was using was limited to 1.5 amps so the transmitter was limited to an output of 38.5 watts on 160 meters. To obtain this output, however, I had to increase the drive to its maximum (16.6 mA dc input). The specified current gain of the 2N5262 is 25 at 1 amp of collector current. Apparently the current gain drops to approximately 10 at 2.5 to

3 amps of collector current. On 80 meters an output of 32 watts required maximum drive. Winding T2 with a 5:1 turns ratio would probably provide enough drive for 50-watts output.

I tried the amplifier on 20 meters and this eventually resulted in two dead transistors. Although the bottom of the collector voltage waveform remains square, ringing due to transformer inductance and stray capacitance continues for most of the time the transistor is off. Although I measured 63% efficiency at about 1-watt output, the amplifier has a tendency to oscillate and to divide frequency. However, I think the circuit would work on 20 meters with different transformers. The series inductance can only be tuned out at the fundamental frequency, and the reactances at the harmonic frequencies make it difficult to obtain the half-sinusoidal current waveforms required in class-D operation.

The variation of output voltage with supply voltage is very linear, and driver feedthrough is only a few milliwatts, so the power amplifier should be suitable for a-m or ssb service (using the EER technique). It might be a good idea to include an automatic phase comparison circuit which would reduce the collector voltage when the output was mistuned (reactive). Such a circuit would be essentially the same as the vswr protection circuits used in early solid-state rf power amplifiers.

modulation

All of the class-D rf amplifiers discussed so far are constant carrier (amplitude) types. If these amplifiers are to be used for a-m or ssb service, they must be modulated. Modulation is most easily accomplished by varying the collector supply voltage with a class-D audio amplifier. For ssb service, this is an application of the envelope elimination and restoration technique discussed earlier. The use of both class-D audio and rf amplifiers produces a highly efficient linear rf amplifier. Collector modulation of the class-D rf amplifier is very linear and there is no need to modulate the driver (in a class-B or -C amplifier, both the power amplifier and the driver must be modulated to prevent overdrive).[14]

Pulse-width modulation can be adapted to produce a modulated rf carrier as shown in **fig. 16**. Because the carrier component of a pulse train is proportional to the sine of the pulse width, rf pulse-width modulation (PWM) must be generated by comparing the desired modulation to a full-wave rectified cosine wave at the carrier frequency. The output amplifier may be either monopolar or bipolar.[13]

Bipolar rf PWM has essentially no splatter due to non-linear modulation of the harmonics of the carrier. Although the rise times required by rf PWM are no higher than those required by a constant-carrier class-D rf amplifier, rf PWM requires switching stages to be controlled as in audio PWM, and this makes it much more difficult to implement. You might be able to build an rf PWM amplifier for operation on 160 meters, but with present devices, it would be difficult to use at much higher frequencies.

vacuum-tube circuits

If you are thinking about a legal limit transmitter, you might consider using vacuum tubes. Power handling capabilities are much greater than those of transistors, and filament power need not be counted as part of the input power. The same configurations apply to tubes as well as to transistors. A vacuum tube can be turned off quite efficiently by driving the grid negative, and this consumes very little power. Driving the grid positive forces the tube to saturate, causing it to act like a resistor. However, driving the grid positive causes grid current to flow and consumes drive power.

What you can obtain with tube switches must be determined by using the characteristic curves of the tubes you plan to use. Generally the saturation losses will be greater than those of transistor amplifiers. However, the output matching networks may be more efficient, and saturation losses occur in class-B vacuum tube amplifiers as well as class-D amplifiers.

When an application calls for a power level or frequency where class-D operation is impractical, there are several other types of amplifiers which have efficiencies better than that of class-B. These include class-C, of course, and a couple of amplifier circuits which don't really operate as class-A, -B, -C *or* -D. Since the envelope elimination and restoration technique makes it possible to use nonlinear rf amplifiers in linear service it is possible to design the rf amplifier for efficiency rather than linearity.

Class-C operation has been used for many years to build rf amplifiers with improved efficiency.[19,20] A class-C amplifier has the same basic circuit (fig. 17) as a class-A or -B linear amplifier, but it is biased and driven so the transistor conducts current for less than half the time. Current is thus drawn through the transistor when the voltage is lowest, resulting in less power consumption.

The reduced conduction angle can be obtained at various combinations of bias and drive, and the current pulse need not be exactly sinusoidal. There are some general rules for class-C operation:

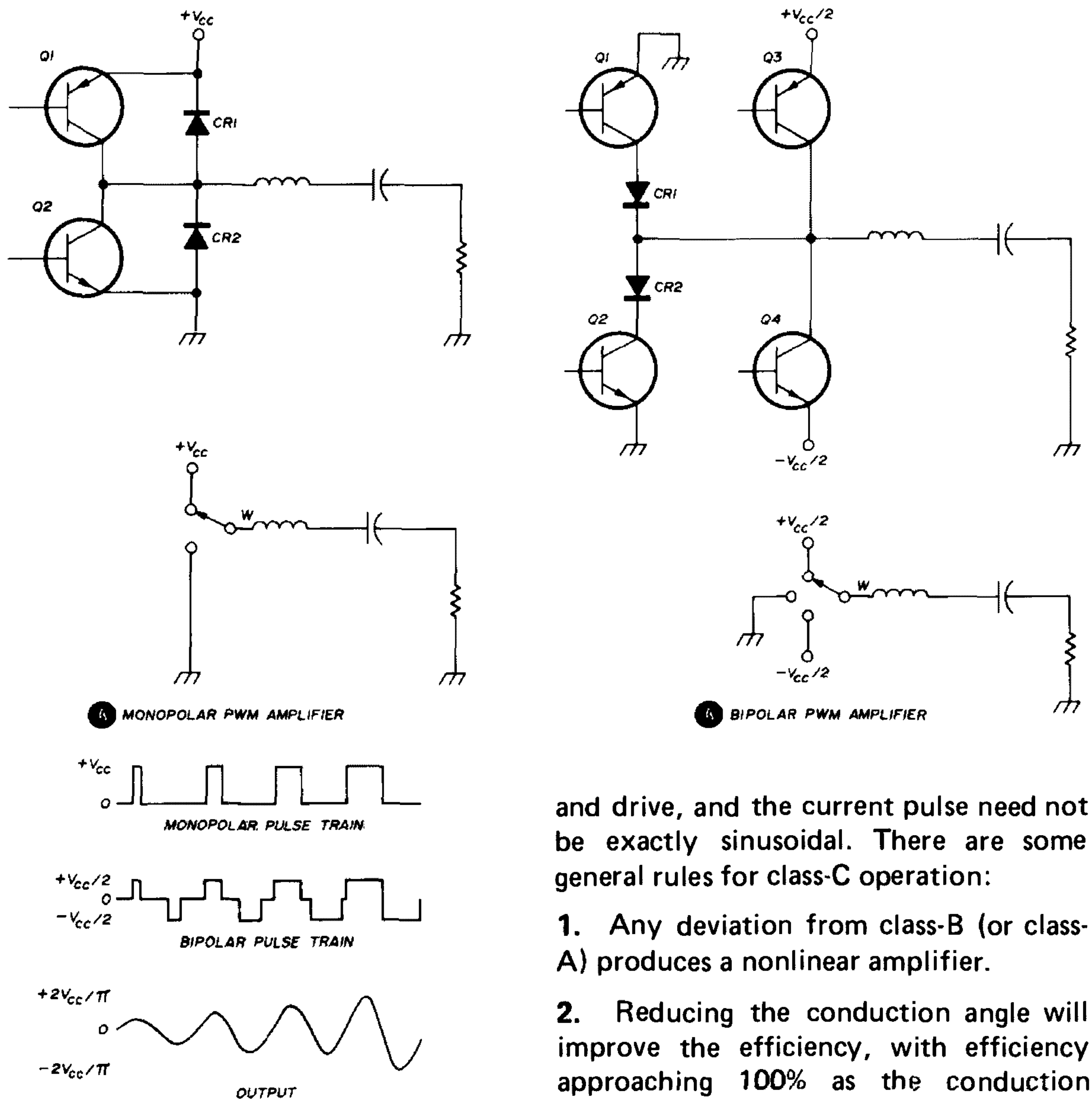


fig. 16. Radio-frequency pulse-width modulation system.

1. Any deviation from class-B (or class-A) produces a nonlinear amplifier.

2. Reducing the conduction angle will improve the efficiency, with efficiency approaching 100% as the conduction angle approaches zero.

3. As the conduction angle is reduced, either output power will decrease or peak collector current will increase.

4. The necessary increase in drive power is approximately inversely proportional to the conduction angle.

These rules will help you to estimate what performance can be obtained by

converting a class-B amplifier to class-C. Class-C can be used to obtain higher efficiency, but it won't be a panacea because drive requirements and peak current will increase with increasing efficiency.

envelope feedback

A technique called envelope feedback can be used to make a nonlinear rf amplifier operate as if it were linear, provided there is some sort of gain control which can be applied.[21] Basically, this technique uses a feedback loop to look at the desired output and the actual output, and correct amplifier gain *so the two are equal.*

Envelope feedback is not a new idea, but today it is much easier to implement due to the abundance of integrated circuits. It is no different from other audio feedback and requires a frequency response from dc to about 10 kHz for ordinary ssb signals. To keep the feedback loop from oscillating, it is necessary that the open-loop gain (point A to point E in fig. 18) be zero for frequencies with 180° or more phase shift. Many books have been written on feedback theory, so I will not go into it any deeper.

Class-C amplifiers are an ideal application of envelope feedback. The bias network normally used to bias a class-B transistor amplifier slightly into conduction can be eliminated and the resulting nonlinearity corrected by feedback. Eliminating the bias network will result in power savings without further reducing the conduction angle. A wide-range gain *control must* be *used ahead of the power* amplifier to reduce the drive. Another possibility is to control the bias of the stage itself; more negative bias means a smaller conduction angle and less output.

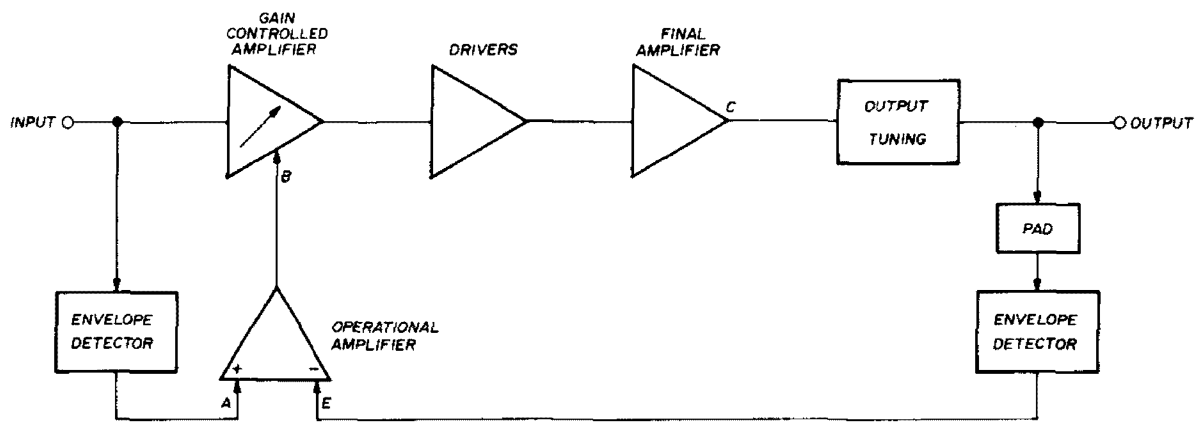


fig. 18. The envelope feedback technique shown here can be used to make a non-linear rf amplifier operate as if it were linear (see text).

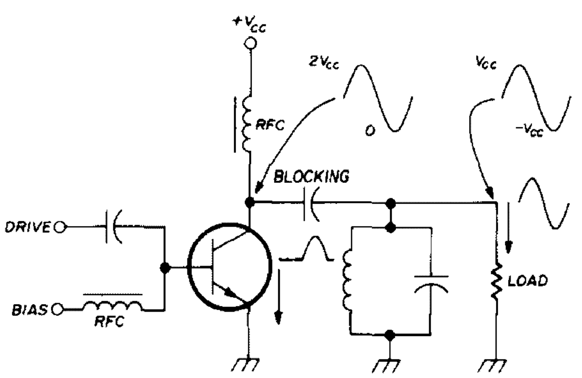


fig. 17. Basic solid-state class-C rf amplifier circuit.

multiple tuning

The multiply-tuned rf power amplifier uses only a single switching transistor and can (ideally) have an efficiency of 100 percent.[22,23] The transistor operates as if it were part of a class-D voltage-switching amplifier. When the transistor is *off*, the current path is provided by a complex tuning network rather than a second transistor. This type of amplifier has been given a different name by practically every author who has written about it; some of the names are: optimally-loaded and overdriven class-B,

multiple-resonator class-C, single-ended class-D, and class-CD. My own name is class-E. This technique is particularly useful at vhf where driving a pair of transistors 180° out of phase becomes difficult.

A voltage-switching class-D amplifier has a *square-wave voltage and half-wave* rectified sine-wave current, arranged so that one or the other is always zero in each transistor. Examination of the waveforms reveals that the voltage waveform contains a fundamental and *odd* harmonics, while the current waveform contains a fundamental and *even* harmonics. To maintain both waveforms with a single transistor requires an output network which is resistive at the carrier frequency, a short-circuit to even harmonics, and an open-circuit to odd harmonics. The series of parallel-tuned circuits in **fig. 19** pass even-harmonic current freely while preventing odd-harmonic current from flowing. The parallel-tuned output circuit passes the even-harmonic current to ground and forces the fundamental frequency current into the load. Ideally, the transistor consumes no power since it always has either zero voltage or zero current.

When this technique is used at lower frequencies, only resonators for the fundamental frequency and third harmonic are usually used, and the circuit is referred to as "third-harmonic peaking." Efficiencies of 90-percent have been reported.

At higher frequencies the series of parallel-tuned traps can be replaced by a quarter-wavelength transmission line as shown in **fig. 20**. The parallel-tuned output circuit provides a short-circuit to all harmonics. The quarter-wave line transforms the short-circuit into an open-circuit for odd harmonics and a short-circuit *for even harmonics. In addition,* it can be used to transform the actual load resistance to the desired load line. If the characteristic impedance of the line is R_o,

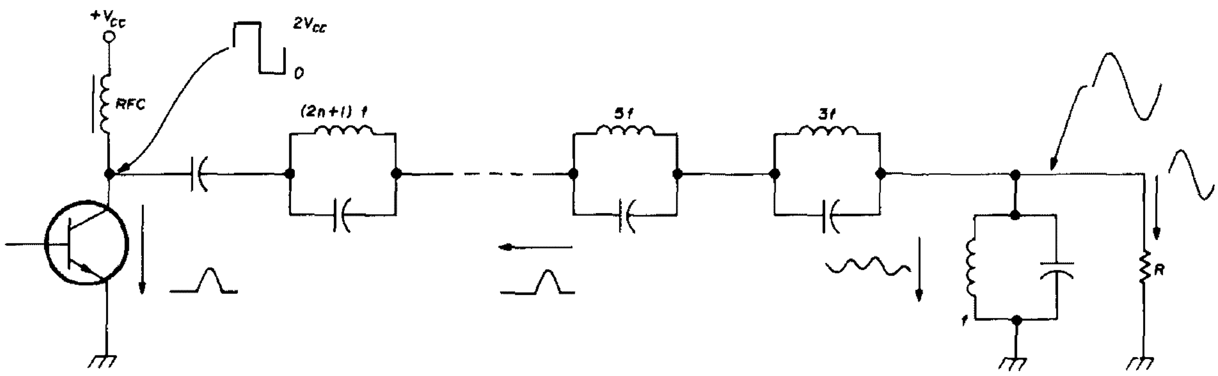


fig. 19. Basic multiple-resonator rf power amplifier. Practical circuit is shown in fig. 20.

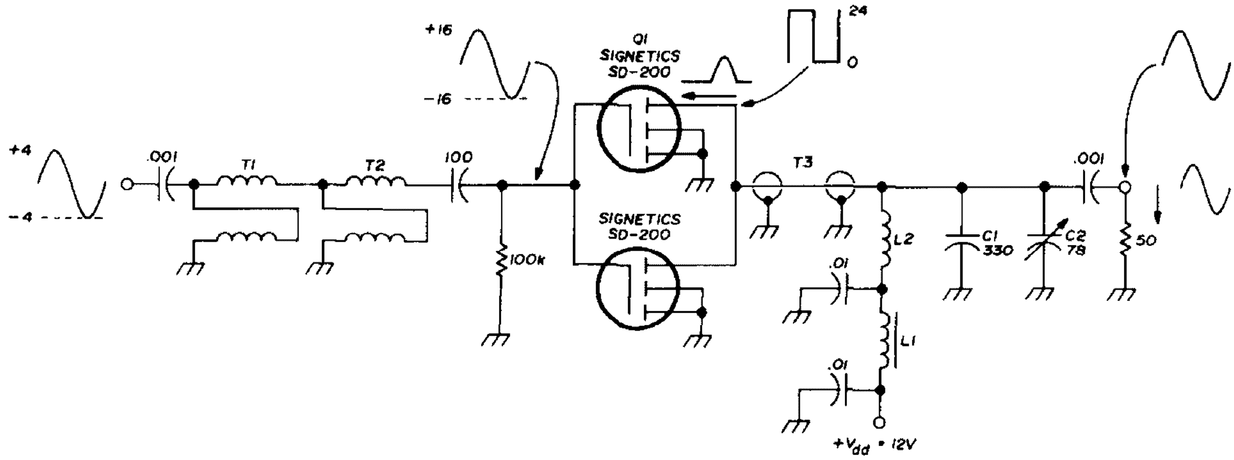
and the actual load resistance is R_L, the load line is

$$R = R_o^2/R_L$$

The circuit shown in **fig. 20** can be adapted to a 300-mW walkie-talkie for 10 or 6 meters (or 2 or 1¼ meters, too, if you're very careful about how you put it together). I tested it at 25 MHz and measured 73% efficiency.* The output circuit has a Q of 3, and the transmission-line transforms the 50-ohm output to a 300-ohm load line.

When saturated, the fet becomes a resistor (about 45 ohms in this case). Since the fet requires negligible drive power, it should be possible to obtain the 16-volt peak sine-wave drive directly from the oscillator tank circuit, eliminating the need for a driver. Remember that matching could also be accomplished with a pi-network—this would allow the use of other transmission-line impedances.

*This circuit was the author's entry in Signetics' D-mos fet contest and will be the subject of a future Signetics Applications Note.



C2	78-pF variable (E.F. Johnson 158-4)	T1,T2	11 turns no. 26 twisted pair on Perma-core 57-2656 or Micrometals T30-6 core
L1	2.2 μH rf choke (Delevan 1025-28)	T3	piece of 125-ohm coaxial cable (RG-63B/U), 112.2" (2.85 meters) long
L2	106 nH (4 turns no. 26 wire on Perma-core 57-2656 or Micrometals T30-6 core)		

fig. 20. This 300-mW multiply-tuned rf amplifier designed for operation on 25 MHz has efficiency of 73%. In this circuit transformer T3 is a section of coaxial transmission line.

capacitor-shunted switch

The capacitor-shunted switching (CSS) amplifier is another type of rf amplifier which uses a single switching transistor and doesn't behave like any of the four standard amplifier classes.[24] I sometimes call it class-F (for far-out). Instead of regarding collector capacitance as an impediment to its operation, the CSS amplifier uses it as an integral part of the circuit.

The shunt capacitor (**fig. 21**) provides a current path when the transistor is *off*. The rf choke acts as a constant dc current source. The series-tuned circuit forces sinusoidal current to flow into the load, and has a high input impedance to harmonic voltages. The difference between the current flowing into the circuit through the rf choke and that flowing in the output either flows through the transistor (when it is *on*) or charges the capacitor (when the transistor is *off*). The fundamental frequency component of the collector voltage is the voltage which appears across the load.

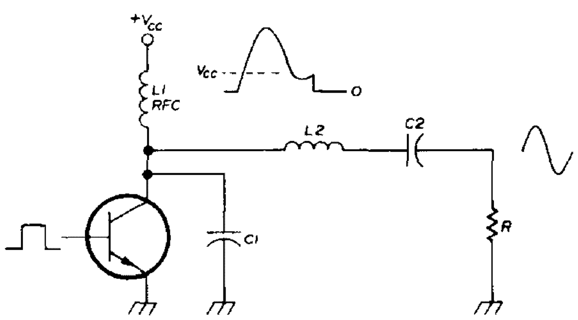


fig. 21. Capacitor-shunted switching (CSS) amplifier can provide operating efficiencies of nearly 100%.

Any energy stored in the capacitor at the time the transistor switches *on* will be dissipated in the transistor. Recently, Sokal discovered that mistuning the resonant circuit so that it is inductive causes the capacitor voltage to drop to zero at the time the transistor switches on, producing 100% efficiency.[25]

Analyzing this amplifier is fairly complicated, but the resulting design equations given by Sokal are *fairly simple:*

$$L2 = \frac{Q\,R}{\omega}$$

$$C2 = \frac{1 + 1/Q}{\omega Q R}$$

$$C1 = C2 \cdot Q/6.3$$

$$P = \frac{2V_{cc}^2}{\left(\frac{\pi^2}{4} + 1 \quad R\right)}$$

where Q is the Q of the output circuit. Adjustment for peak efficiency may be

somewhat critical, and the amplifier will have inefficiency due to saturation voltage and rise time, just as a class-D amplifier does. However, the simplified output circuit (compared to multiple tuning) and reduced driving problems (compared to class-D) may be worth it, especially at vhf.

summary

In this article I have tried to give an overview of high-efficiency amplifiers, particularly class-D amplifiers, which might be useful to amateurs. There are a lot of variations beyond those mentioned here. Understanding how they work requires careful attention to voltage and current relationships and a little algebra. With a little luck, an enterprising amateur should be able to implement some of these ideas into his own equipment. I would like to hear from anyone who does.

I would like to thank my friends Rich Grimsley, WA3JEL, and Bob Calderwood, WA7ANT, for their very helpful comments on the rough draft of this manuscript.

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ham radio

feed system

for log-periodic antennas

How to find the optimum point for feeding a multi-band log periodic

G.E. Smith, W4AEO, 1816 Brevard Place, Camden, South Carolina 29020

Log-periodic antenna texts provide little information on methods of getting a practical transmission line matched to the antenna feedpoint. References 2, 5 and 7 describe methods of matching and feeding vhf log periodics by use of boom balun to coax matches, but most of these systems are not suitable for high-frequency log periodics.

The original log-periodic antennas built here[9] were first fed by using a 1:1 balun with the balanced winding across the normal short-element feedpoint. Although the swr was relatively low across the 20-meter band and not too bad on 15, there was some variation between these two bands. On 10 meters there were some bad swr excursions when going from 28.0 to 29.7 MHz, some exceeding 2.5:1, showing a bad mismatch on these higher frequencies. Equipment was not readily available to make a complete swr run over the antenna's entire bandwidth, 14 to 30 MHz.

Though some of the vhf references [1,2,3,etc.] indicate that the swr could be expected to go up to 2.5:1 over a log periodic's bandwidth, it was felt this could be improved by a better matching system between the transmission line and the antenna. Upon checking several log periodics at the normal short-element feedpoint using the Omega Antenna

Noise Bridge, it was noted that there was considerable variation in impedance over the three bands.

analysis

Upon anaylzing this result it became evident that the active or driven elements (one-half wavelength long at a specific frequency) were at various electrical distances (1/4 wave-wise) from the feedpoint. In **fig. 1** for example, the second element, which is the driven element on 20 meters, is 35 feet (10.7 meters) or approximately a half wavelength to the rear of the feedpoint at the front of the antenna. Since previous tests with the antenna bridge had indicated the driven element exhibited 30 to 33 ohms input impedance, it can be assumed that this impedance would be repeated at the feedpoint since the two are connected with a feedline *which is a half-wavelength* long. This was confirmed by tests with the bridge on 20 meters.

Consider the situation on 15 meters. Element 6 is a half-wavelength long on that band so it becomes the active or driven element. It is located about 16.5 feet (5.1 meters) from the feedpoint, but a quarter-wave of open-wire feedline on 15 is about 11.25 feet (3.4 meters) and a half-wave about twice that long (22.5 feet or 6.9 meters). Thus, the feedpoint is about halfway between a high- and low-impedance point on 15 meters.

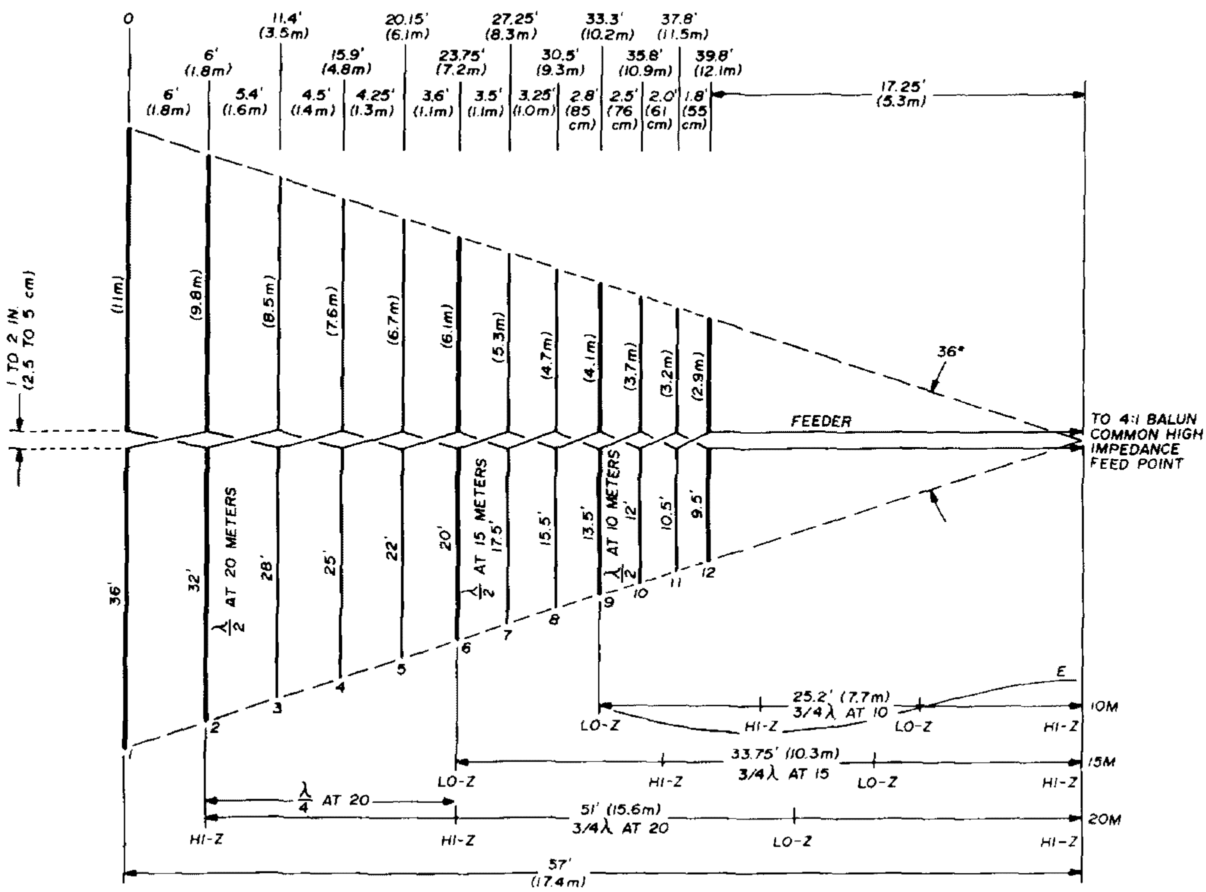

fig. 1. A 12-element three-band log-periodic antenna showing the electrical relationship of the extended feedpoint to the driven element on each of the three bands.

Looking at 10 meters, element 9 is about 16 feet (5 meters) long so it is the active element on that band. It is about 6.5 feet (2 meters) to the rear of the feedpoint. A quarter-wavelength feedline on 10 meters is about 8.4 feet (2.6 meters) and a half wave, 16.8 feet (5.2

meters). Again, the feedpoint is at an intermediate point with respect to the active element.

In summary, it will be noted that although the feedpoint at element 12 presents a fairly predictable impedance on 20 meters, it presents a highly variable match on 15 and 10. This was confirmed by swr readings across each of the three bands.

wavelength open-wire line acts as a matching transformer.

The low and high impedance points along the open-wire line for each of the three bands are shown in fig. 1. Incorporation of this modification extends the feedpoint 17.25 feet (5.3 meters) forward of the short-element end of the antenna.

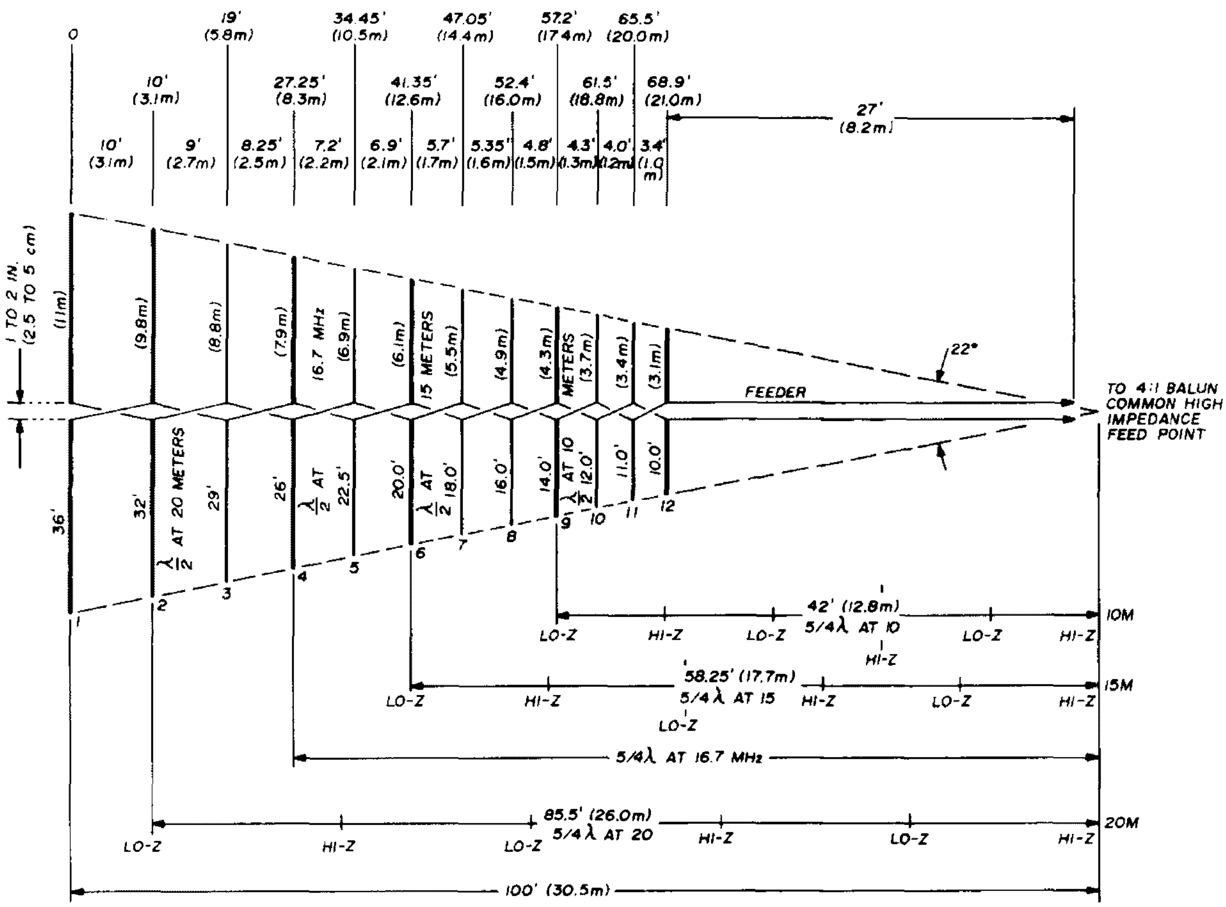

fig. 2. An extended 12-element three-band log periodic with a 5/8-wavelength extended-feed matching system.

moving the feedpoint

By extending the open-wire feedline toward the antenna apex as shown in fig. 1, a point is reached near the apex where elements 2, 6 and 9 are all about 3/4 wavelength from this common, higher impedance, feedpoint for their respective bands. The impedance at this point appears to be on the order of 200 to 300 ohms, so it can be fed with 50-ohm coax through a 4:1 balun with a satisfactory match on all three bands. The 3/4-

In **fig. 2** the same principle of an extended open-wire feeder is applied to matching a longer log periodic with a 70-foot (21.4-meter) boom length and 22-degree apex angle. This requires an open-wire feeder 5/8-wavelength long to reach the common-impedance feedpoint, also approximately at the apex angle. In this case the open-wire feeder has been extended 27 feet (8.2 meters) from the center of the short element. Note that these extended feeders can hang down from the short-element end of the antenna if necessary; they need not be extended horizontally as shown.

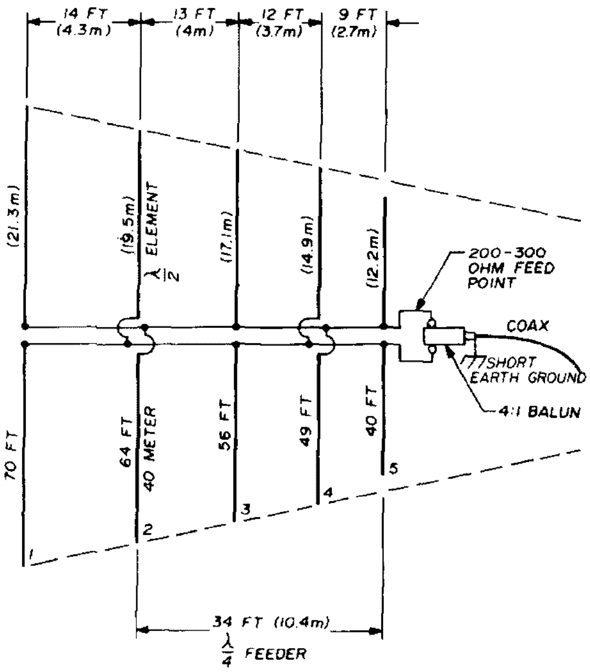

fig. 3. A 5-element, 40-meter monoband log periodic. The center of the short element is just a quarter wavelength from the driven element, providing an optimum feed point.

The longest 20-, 15-, and 10-meter log periodic in use here (17 elements, with a 100-foot (30.5-meter) boom and 16 degree apex angle) requires seven quarter-wavelengths between each active element and the common feedpoint.[14] Further, it will be noted that any active element on any of the antennas shown is an odd number of quarter wavelengths, three in **fig. 1** or five in **fig. 2**, from the common feedpoint. For example, element 4 in **fig. 2** is 28 feet (8.5 meters) long and will resonate at 16.7 MHz. It is approximately 5/4-wavelength from the common feedpoint at this frequency, so we can therefore assume that this antenna is a true log periodic.

The monoband log periodics of **fig. 3**, tested here on 10, 15, 20 and 40 meters in 5-element versions (some 4- and 6-element types were also tested) have all had element 2 exactly a quarter wavelength from the high-impedance, short-element feedpoint. These have worked extremely well using a 4:1 balun to match them to coaxial transmission lines. The swr has been relatively low across each band.

Similarly, the single-band vertical monopole log periodics of **fig. 4**, tested on 40 and 80, also used the quarter-wavelength feed and were similarly flat.[15] The swr readings on the 80-meter version were:

4.0 MHz	1.25:1
3.9 MHz	1.4:1
3.8 MHz	1.2:1
3.7 MHz	1.1:1
3.6 MHz	1.2:1
3.5 MHz	1.2:1

similar approach

This system for feeding log-periodic antennas was believed to be original at

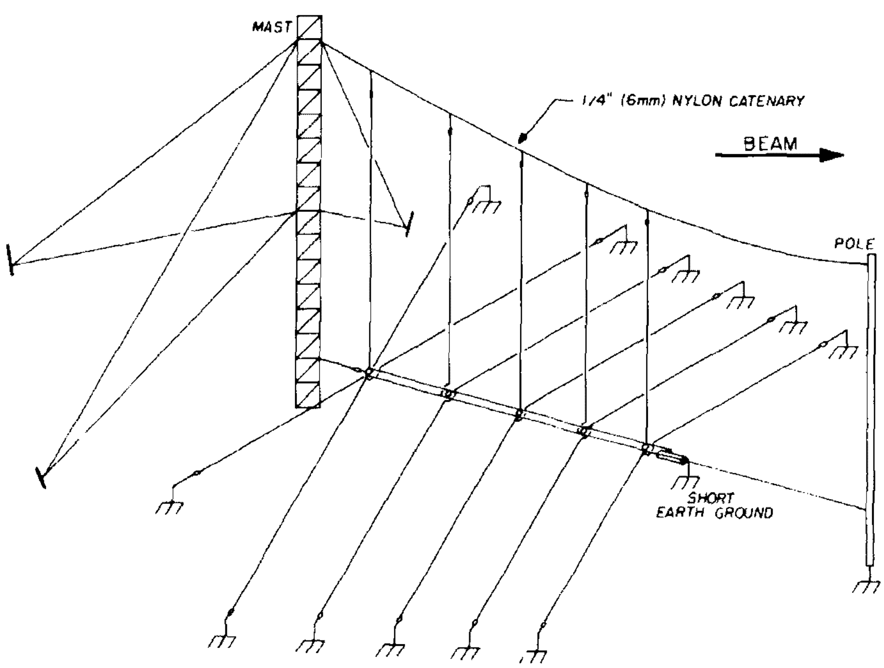

fig. 4. Overall view of the five-element vertical monopole log periodic. Construction details and dimensions are shown in fig. 5 and table 1.

table 1. Vertical monopole log-periodic antenna dimensions (5-element arrays).

	frequency in MHz			
element	3.5-4.0[1]	3.8-4.0[2]	7.0-7.3[2]	14.0-14.35[1]
E1	70' (21.4 m)	65' (19.8 m)	35' (10.7 m)	17.5' (5.3 m)
E2	67' (20.4 m)	62' (18.9 m)	33' (10.0 m)	16.5' (5.0 m)
E3	58' (17.7 m)	55' (16.8 m)	28' (8.5 m)	14.0' (4.3 m)
E4	50' (15.3 m)	45' (13.7 m)	24.5' (7.5 m)	12.2' (3.7 m)
E5	43' (13.1 m)	40' (12.2 m)	20' (6.1 m)	10.0' (3.0 m)
S1	30' (9.2 m)	26' (7.9 m)	14' (4.3 m)	7.0' (2.1 m)
S2	27' (8.2 m)	24' (7.3 m)	13' (4.0 m)	6.5' (2.0 m)
S3	24' (7.3 m)	23' (7.0 m)	12' (3.7 m)	6.0' (1.8 m)
S4	19' (5.8 m)	18' (5.5 m)	9' (2.7 m)	4.5' (1.4 m)
total length	100' (30.5 m)	91' (27.8 m)	48' (14.6 m)	24' (7.3 m)
mast height	80' (24.4 m)	75' (22.9 m)	50' (15.3 m)	30' (9.2 m)
pole height	45' (13.7 m)	40' (12.2 m)	25' (7.6 m)	20' (6.1 m)

1. Calculated design, not actually built and tested.
2. Built and tested design, with measured swr under 1.5:1 over frequency range shown.

the time I worked it out. However, after it was mentioned briefly in a previous article[9] it was learned that Ray Rosenberry, K8EBF, developed a similar method that was described and covered by his patent of 16 February, 1971, which covers "Broad Band Transformer Antenna and Related Feed System."[13] Therefore, I do not claim the odd quarter-wavelength feed method for log periodics to be original. K8EBF and I have since exchanged several letters regarding these log-periodic feed methods.

In any case, it is hoped the above information on improved methods of feeding log-periodic antennas will be helpful to amateurs using these interesting antennas. I would like to hear from anyone trying this technique.

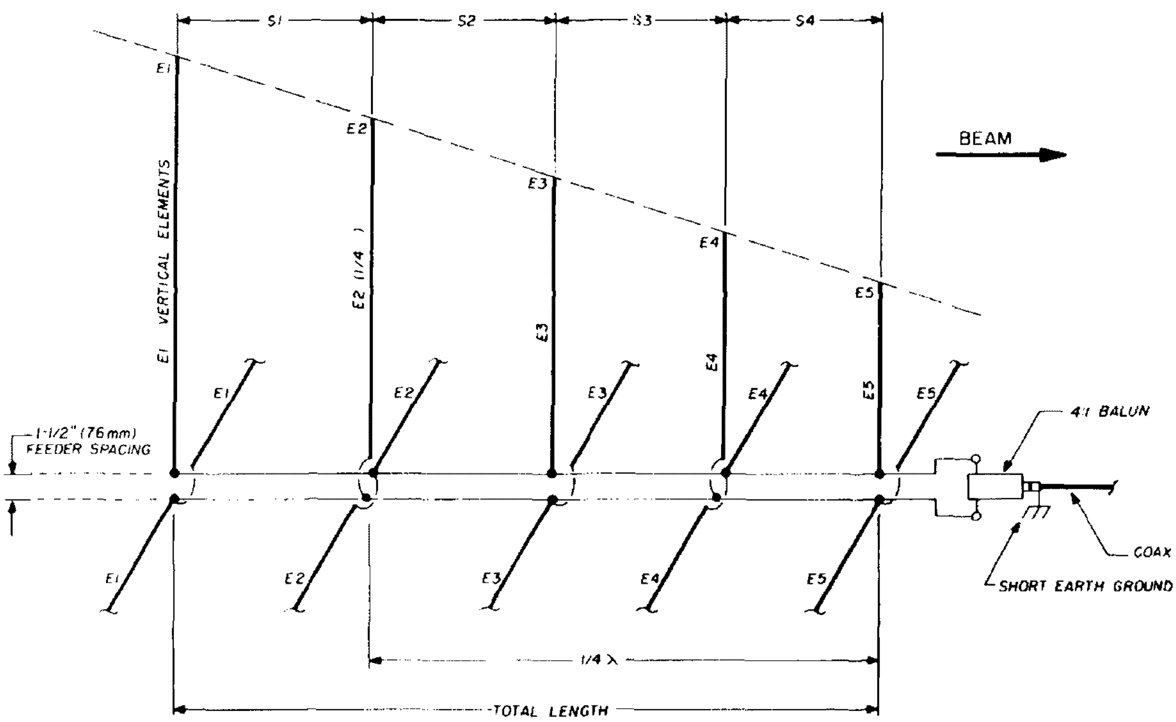

fig. 5. Construction details for a 5-element vertical monopole log periodic. Dimensions for 80, 75, 40 and 20 meters are given in table 1.

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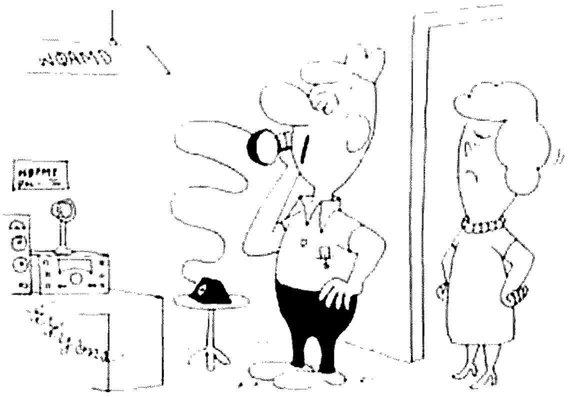


"Ever since I bought that $450 receiver, I've had a communication problem — my XYL!"

high-stability crystal oscillator

The Goral oscillator — a new high-performance crystal-oscillator circuit that features excellent load capacitance correlation and temperature stability

Donald L. Stoner, W6TNS, Director of Marketing, SGC, Inc.*

One of the **most frustrating** problems faced by a newcomer to the crystal manufacturing business is that of capacitance correlation of the oscillators. Although 32 pF has been the recommended standard for many years, few engineers seem inclined to use this value in their new circuits. In many cases the engineer gets the circuit working to his satisfaction and is reluctant to change any of the component values. Usually, the burden of correlation falls on the crystal supplier.

A second problem, which is no less irritating, is that of activity requirements. The crystal vendor may supply perfectly good crystals, only to find that many will not oscillate in the user's circuit.

While the crystal oscillator circuit to be described will not cure any of these past problems, it can certainly minimize them in the future. Although the oscillator configuration (which is an adaption of the basic Colpitts circuit) may not be new, I am unaware of any similar designs. It was developed independently in the quartz crystal laboratory at SGC, Inc., and has the following features:

1. The ability to correlate at virtually any capacitance between **5** and **50** pF. In fact, the crystal will oscillate very close to its series-resonance point in this circuit.

2. The ability to make crystals oscillate with equivalent series resistances (ESR) well above the Mil-spec maximum. The circuit is sufficiently active to oscillate quartz blanks which have not been plated during the manufacturing process.

3. The temperature stability is essentially independent of the solid-state devices used. Capacitance variations due to the semiconductor junctions are virtually swamped out due to the large shunt capacitance in the circuit.

4. The circuit is easily adapted to "rubber" the crystal with semiconductor variable-capacitance diodes and exhibits a very wide pulling range due to its high activity.

Colpitts oscillator

The simplified circuit for the transistorized Colpitts oscillator is shown in **fig. 1**. Note that the rf ground has been shifted from the customary position at the collector. Oscillation occurs due to the **180°** phase-shift through the transistor and the additional **180°** inversion across the tuned circuit represented by the crystal. The capacitor values are determined by the circuit impedance and frequency range.

The basic circuit for the fet Colpitts oscillator is quite similar and is shown in fig. **2**. Note that the rf ground has been moved to its customary position by bypassing the source (collector), leaving the drain (emitter) at an rf potential above

*SGC, Inc., Frequency Control Division, 13737 SE 26th Street, Bellevue, Washington 98005.

ground. Thus, one end of all crystals in a bandswitching circuit can be grounded. The fet Colpitts circuit is also very useful because the temperature characteristics are excellent. The fet causes virtually no drift with temperature.

Fig. 3 shows a practical application for the fet Colpitts oscillator. The values have been optimized for the 10-to-20-MHz range. Note that this circuit also incorporates a leveling diode, CR1, that rectifies the rf voltage across the crystal. Thus, the more active the crystal, the more negative bias which is produced. This, in turn, reduces the gain of the oscillator. Thus high and low activity crystals tend to produce the same output voltage at the drain of the fet.

Unfortunately, the fet Colpitts is not a particularly active circuit and seems to require lower resistance crystals than its transistor counterpart. Crystals with an ESR above 20 ohms appear to be quite

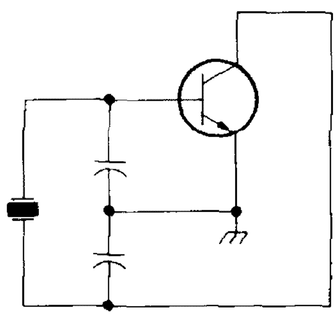


fig. 1. Basic Colpitts oscillator using an npn transistor.

sluggish. In this instance, the ESR referred to is measured at parallel resonance. This is the mode the crystals oscillate on in this circuit.

The sluggishness of the fet circuit is due primarily to the shunt loading placed across the crystal by the netting capacitor, C1, and the diode, CR1. If a variable-capacitance diode is used in the circuit (in addition to the netting capacitor), crystals with ESRs of less than 16 ohms are required for reliable starting.

Goral oscillator

Obviously, it is possible to make the basic fet Colpitts oscillator more active by operating it at higher drive levels. However, for best stability, the crystal drive level should be kept as low as possible, consistent with reliable starting. If the oscillator loop gain can be increased without increasing crystal current or substantially increasing drive level, the performance of

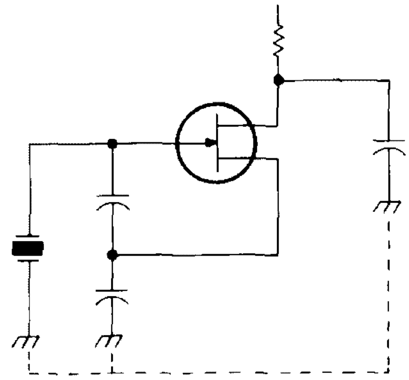


fig. 2. Basic Colpitts oscillator circuit using an fet.

the basic Colpitts fet oscillator can be noticeably improved.

This is what is done in the Goral oscillator circuit shown in **fig. 4**. While only a few components are added to the basic circuit of **fig. 3**, and the circuit changes are subtle, the difference in performance is extraordinary. Again, the component values are optimized for 10 to 20 MHz. In this circuit transistor Q2 acts as an emitter follower to provide power gain for the feedback energy without changing the phase angle of the signal. The increased feedback permits increased values of C2 and C3 from those shown in **fig. 3**. This further increases the tempera-

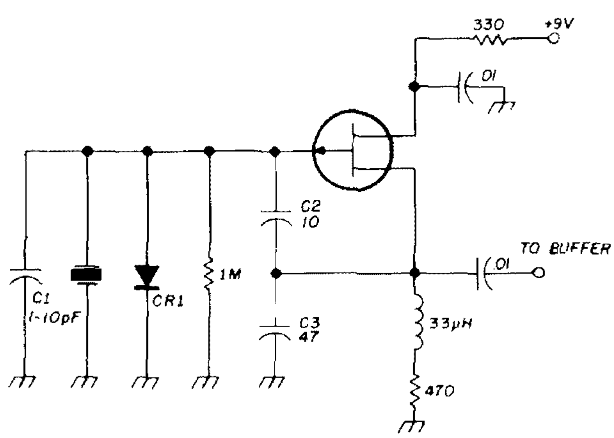


fig. 3. Practical Colpitts oscillator using an fet. Component values have been optimized for the 10- to 20-MHz frequency range.

ture stability of the circuit. Transistor Q3 is a simple buffer to prevent oscillator loading.

Don't be concerned by the strange numbers associated with the transistors. Device Q1 is a Motorola jfet and was selected because of its low cost ($.38 each). Practically any junction-type field-

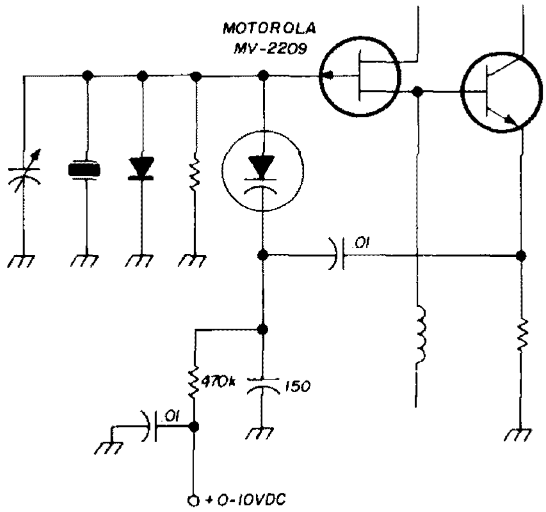


fig. 5. A varicap may be used in the Goral oscillator, as shown here, to vary the output frequency from a variable dc source.

effect transistor will work in the circuit without affecting performance. The MPS-5172 may also be an unusual number. Again, it is the lowest cost Motorola rf small-signal device in their line ($.21 each). There are HEP equivalents, but, frankly, they are more expensive because

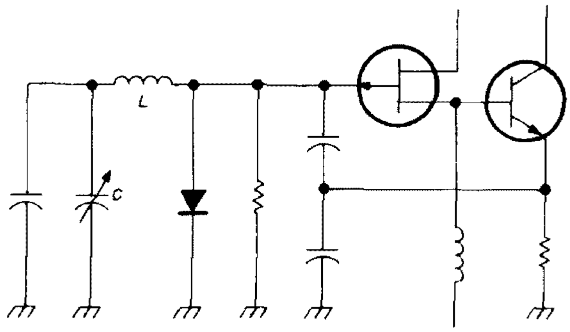


fig. 6. Using the Goral oscillator as a highly stable vfo. The L/C values are chosen for the desired output frequency.

of the packaging. Again, transistors Q2 and Q3 are totally non-critical and virtually any general purpose npn rf type should work equally well. The agc diode, CR1, is a 1N914 or 1N4148.

Fig. 5 shows a simple variation to the basic Goral oscillator which incorporates a variable-capacitance diode. This allows the frequency to be varied with an external dc source (such as a clarifier on fixed-frequency ssb equipment) or modulated for fm applications.

Although the circuit of fig. 6 is untested, it should make an outstanding vfo. Any temperature instability and drift which occurs in this circuit is the result of the tank circuit and not the devices or components.

A printed-circuit board for a simple crystal test oscillator is shown in fig. 7. Note the only variation from fig. 4 is the

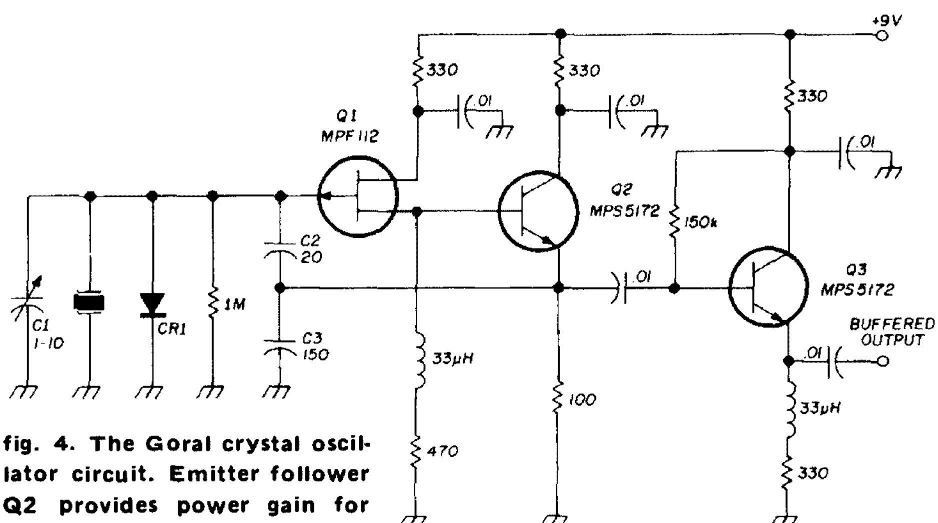


fig. 4. The Goral crystal oscillator circuit. Emitter follower Q2 provides power gain for the feedback energy, providing high crystal activity without affecting the crystal drive level. Transistor Q3 is the output buffer. Diode CR1 is a 1N914 or 1N4148.

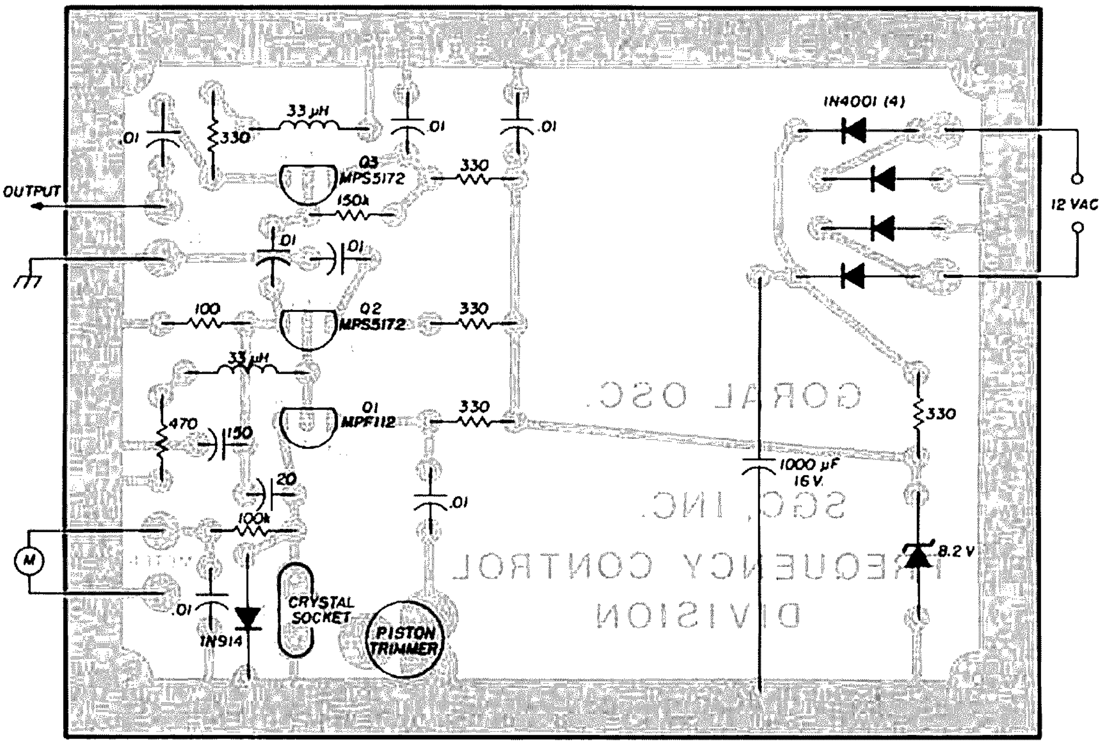


fig. 7. Printed-circuit component layout for a simple crystal test set using the Goral oscillator. This same circuit board may be used for the circuit shown in fig. 4 (see text). Full-size PC board is shown in fig. 8.

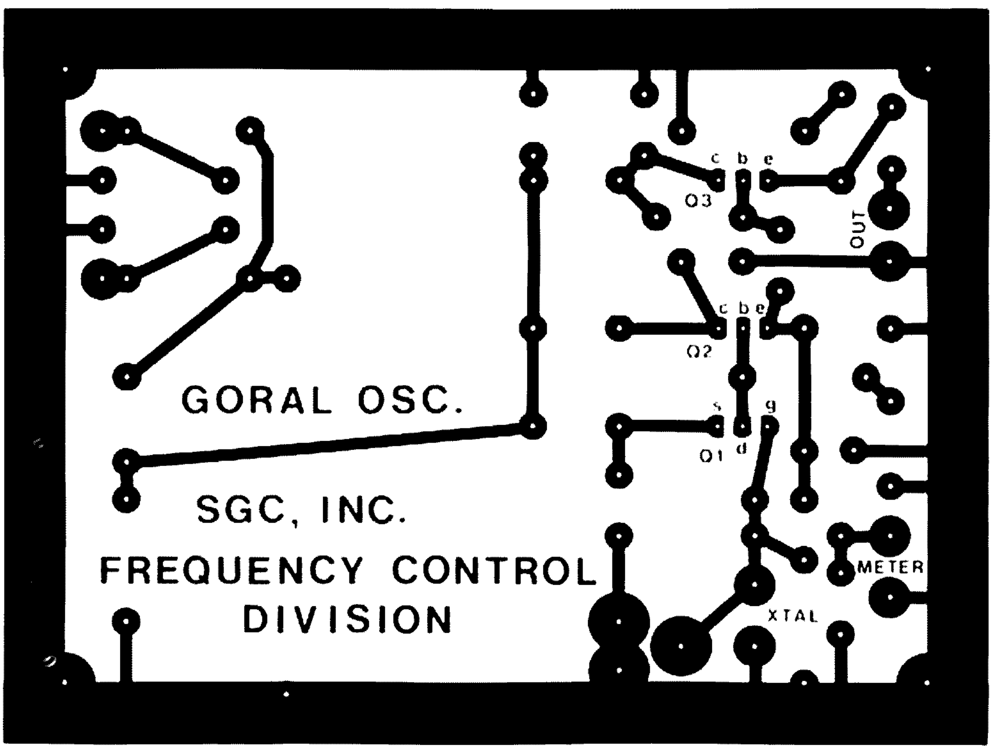


fig. 8. Full-size printed circuit board for a simple crystal tester.

value of the gate resistor for Q1. Here it is reduced to 100k to permit reading the gate current (and, therefore, crystal activity) on the meter.

I wish to thank Pierre Goral, who developed the circuit, for his assistance in preparing this article.

ham radio

simple tunable receiver modification for vhf fm

Simple vfo adds all-channel receiving capability to two-meter fm transceivers

Paul Franson, WB6VKY, 577 36th Street, Manhattan Beach, California 90266

Vhf fm **transceivers** have become very popular for local communication, taking much pressure off the high-frequency bands, but the typical fm transceiver, with its limited number of channels, lacks the versatility that many hams desire. In Southern California, for example, every two-meter channel is occupied – some many times over – and trying to provide operating capability on each can be very expensive, whether the approach taken is multiple crystals or a frequency synthesizer. Yet most hams would like to operate, or at least have the capability to operate, on many channels. After operating for a while, mostly mobile, on a few channels that were already installed in the used rig I bought, I came to the conclusion that what I really needed was tunable receiving capability, not vast numbers of full channels. A tunable receiver permits me to listen in on other channels, perhaps finding one I'd like to try, or quickly eliminating others because of the geographical coverage, use of the repeater, or even the people on the channel. It's obvious, I believe, that some repeaters welcome new blood, whereas others are occupied by users who prefer to talk to the same people, just as is true in any social organization.

A tunable receiver also permits tuning the input of repeaters when malicious or accidental interference disrupts the frequency, making a little transmitter hunting desirable.

I developed the simple variable-frequency oscillator shown in **fig. 1** to give me the tunable receiving capability needed for the relevant upper two megahertz of the two-meter band. It replaces the first crystal oscillator in an fm transceiver, and tunes over the required range, generally one-third of the first injection frequency; *i.e., it* operates at approximately 45 MHz for popular fm transceivers, a frequency that can easily be changed if needed.

The vfo can either be built into an existing transmitter, or into an add-on enclosure. The circuit itself is very small, and the potentiometer used to tune it can be placed remotely.

circuit description

The vfo uses a field-effect-transistor oscillator *tuned* by *a tuning diode* (variously known as a Varicap, varactor or variable-capacitance diode). A source-follower output stage buffers the oscillator, and the supply voltage is regulated by a zener diode. Stability is certainly adequate for the usual fm receiver, but I haven't checked it on one with a very narrow bandpass.

None of the parts in the vfo seem exotic, and it should be possible for any active building ham to put it together in a few hours out of parts in his collection.

The oscillator uses an inexpensive plastic field-effect transistor in a Colpitts circuit. The transistor is a member of the popular and large Motorola family whose Grandfather is the MPF102, but I recommend that you use one of the better specified versions rather than the MPF102, which has a very wide I_{DSS} range. I used the MPF102, but ended up having to try different values of source resistance to get satisfactory results.

The oscillator tank coil is a simple four-turn coil of heavy silver-plated wire. The silver is obviously not needed, but I found it easier to unwind a short piece of World War II vintage tank-transmitter coil than find tinned number-16 copper wire. A toroidal coil would also be suitable,

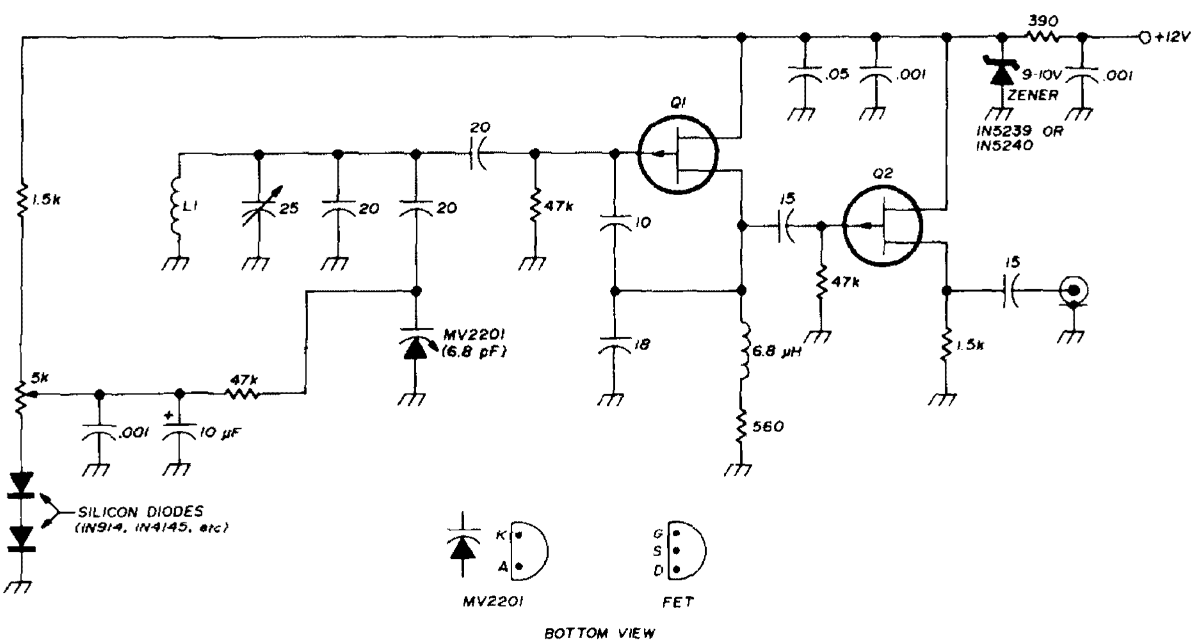


fig. 1. Schematic diagram of the receiving vfo for two-meter vhf-fm equipment. Fets Q1 and Q2 are MPF102 types (2N5668, 2N5669 are better). L1 is 4 turns no. 16, ½" (13 mm) in diameter.

perhaps even necessary, for lower-frequency versions if you want to keep the vfo small.

The circuit is tuned by an inexpensive Motorola tuning diode in a plastic TO-92, D-shaped case like that used for the fet oscillator. The particular diode used, the MV2201, has a nominal capacitance of 6.8 picofarads, but any similar diode — even an expensive military type — is suitable. I used Motorola parts because they are widely available; the HEP versions are quite suitable with the sole warning that their HEP802 fet appears to be like the MPF102, so the same warning about the source resistor applies. If it doesn't work, try another value of resistance.

The varactor tuning makes it simple to tune the unit remotely, and to pick restricted ranges, but a small trimmer capacitor could also be used. In this case, construction is more complicated and critical.

With a variable resistor for tuning, however, you can take a number of approaches. A simple 270° potentiometer will cover the band in one swing, but makes tuning a little tricky. One is easy to add to the front panel of a transceiver, however, and if the tuning range is restricted to part of the band, this should work. Use a composition pot, not a wirewound one, which tends to give step-wise tuning, and the steps aren't likely to be the ones you want.

For best control and calibration, I recommend a ten-turn potentiometer and appropriate digital dial. With a 2000-kHz range, and close to 1000 divisions, repeatability and calibration is excellent. It won't be linear calibration, of course, though if you're a masochist, you could probably come close with proper selection of component values and padding. A wire-wound pot is fine for this use since the steps are much smaller.

It's also possible to use a switch and multiple resistive trimming pots for selecting specific channels. I still recommend that you have provision for variable tuning, however.

You might notice the two diodes at the ground end of the potentiometer. These forward-biased silicon diodes raise the cold end of the pot about 1.4 volts above ground, ensuring that the varactor is reverse biased at any setting of the pot. They also help select the proper bandspread (with the 1500-ohm resistor and 5000-ohm pot), and provide some tem-

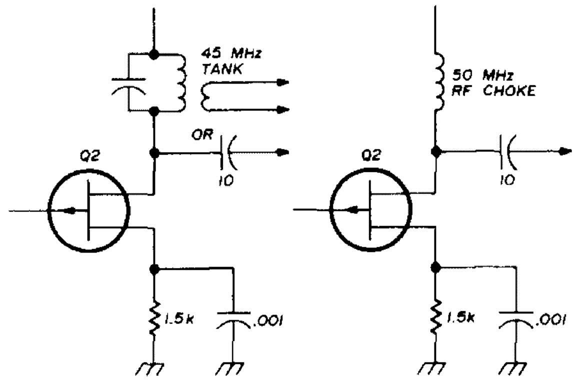


fig. 2. Suggested tuned circuits for obtaining higher output drive from the buffer stage.

perature compensation for the varactor, at least in theory. I don't have the facilities for checking the latter, and am not ambitious or curious enough to find them.

I highly recommend that you use quality mica capacitors in the frequency-determining tank circuit (and that includes all small value capacitors in the circuit except the 15-pF output capacitor). An unstable ceramic capacitor can be very frustrating, particularly if you use the unit in an unheated garage or automobile (admittedly not as big a problem in coastal Southern California as it is in Wisconsin or New Hampshire). The small rf choke, by the way, is not critical in value. The circuit works without it, but exhibits considerable amplitude variation across the tuning range.

The buffer output stage is straightforward, and uses the same type of transistor as the oscillator. The source follower configuration provides adequate drive for popular transistor rigs; if drive is inadequate, it might be necessary to

convert this stage to a tuned-drain version, as shown in fig. 2 (I haven't tried this, though). Even a broadly tuned rf choke might be adequate, but don't bother unless you try. Most quality rigs designed for fm don't require much drive; it would make the crystals drift. A converted military receiver might be another matter.

The vfo operates from about 12 volts, with a 9- or 10-volt zener diode regulating the voltage to the transistors and tuning diode. Little current is required, but if you operate from small batteries, you can dispense with the zener and operate directly from 9 volts or so. The short-term variations should be small enough for the intended use. That's not true in a car, however. The supply voltage there often varies from 11.5 to 14.5 volts.

construction

The easiest way to make the vfo is probably to use a small peice of perforated Vector board. An etched circuit board is prettier, but a lot of trouble if you're just making one. Do mount everything solidly, as in any vfo. I found fairly extensive filtering and shielding necessary, but this might not be critical out in the provinces. Before I shielded the unit, it received amateurs, Adam 12, aircraft and Alice Cooper, simultaneously. A shielded cable is vital for both remote control and output leads, also.

installation

The electrical installation is relatively straightforward; I'll leave the mechanical engineering to you. Simply unplug one crystal, and connect the output of the vfo to the hot end. In the Tempo FMP I use, the high end of the crystal is switched, and the low end is grounded through a small trimmer capacitor. Make sure that you connect the vfo and cable ground directly to the transceiver ground, not through this trimmer. Use the shortest lead possible, and a direct ground.

I recommend Teflon-insulated miniature cable, not RG-174/U with its soft plastic insulation. Most of the fm rigs are tiny, and burning plastic insulation looks and smells horrible. By the way, the battery compartment of my Tempo FMP has space for the vfo.

One warning: It is possible for the crystal oscillator in the rig to take off when you attach the input lead to it, but this can be prevented by installing a small rf choke in the lead from the vfo right at the crystal socket. A couple of small ferrite beads may do the job. If worst comes to worst, you might even have to ground the emitter of the crystal oscillator transistor through a bypass capacitor of, say, 470 pF, when using the tunable vfo. I didn't have any trouble with one vfo I built, but the other needed the rf choke.

calibration

I suggest that you make sure the vfo is working properly before you install it, as the installation will undoubtably be a little tedious if you put it in a transceiver. The easiest way to check operation and range is with a suitable frequency counter. You can also use a monitor receiver, or a signal generator plus the receiver, but these can be confusing due to the numerous images possible. You'll need to touch up the tuning after installation. If the tuning range isn't correct, you can juggle component values. This technique is undoubtably all too familiar to anyone who reads *ham radio* and has read this far.

final note

Just out of curiosity, I replaced the source-follower output stage with a tuned tank circuit on two meters (as in **fig. 2**), then introduced a small amount of audio into the varactor from a dynamic microphone. The combination transmitted good quality fm for a few blocks, but I can't really recommend the technique for general use, not with the sharp receivers (and tongues) found in most areas. I don't even recommend it as a crystal replacement for transmitters for the same reasons.

ham radio

mechanical design of cubical quad antennas

Construction notes for a quad designed to withstand the fury of Canadian winters

R.C. Golding, VE3II, 69 Gordon Road, Willowdale, Ontario, Canada

After using a Yagi antenna for ten years with excellent results, the time came when it had to be replaced. I decided this was a good time to test the relative merits of the quad antenna. The following piece presents the results of four years of work dealing with construction problems of a cubical quad in a Canadian climate. Emphasis is placed on mechanical details such as clips and fasteners, joint interfaces, protection of wire elements and spreader design. On-the-air results are also reported, which are based on qualitative observations.

Electrically, the design is the same as that described by Lee Bergren in the May, 1963, issue of *QST*.[1] The only claim I have to this design is the mechanical features described here.

project objectives

The quad has a bad reputation in Canada because of mechanical problems that occur during the severe winters. My first two models failed during successive winters; the third model survived three Toronto winters with no maintenance. It was designed with the following objectives:

1. Stronger joints between crossarms and boom.
2. Better clamps between spreaders and wire elements.

3. A means to avoid element breakage.

4. A better joint between spreaders and crossarms.

description

The boom is made of 3-inch (7.6-cm) aluminum irrigation pipe instead of the 4-inch (10.2-cm) diameter pipe used in Bergren's design and proved to be quite satisfactory. It is lighter in weight and crossarm-to-boom joints remained tight by minimizing sway in the spreaders, which could elongate the holes in the boom. In a commercially-built quad, the holes through the boom have additional bearing surfaces provided by sleeves shrunk onto the boom. This is difficult for the amateur to achieve, so it was necessary to make a tight fit with the materials available. If you have the means to shrink sleeves onto the boom at these points, the advantage is obvious. Aluminum irrigation pipe is not always perfectly circular, and this in itself presents a difficulty when making sleeves or spider clamps.

drilling the boom

As explained in the article by Bergren, holes are drilled at 90 degrees through the

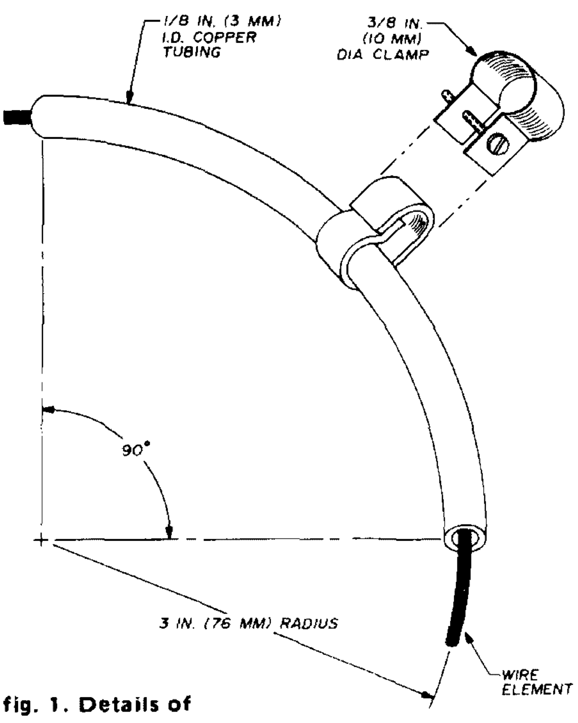


fig. 1. Details of the clamp and support for antenna elements.

boom 1/8 inch (3 mm) apart. These holes are to accommodate the square aluminum plate used as a support. Certain precautions must be observed, however, in drilling the holes to make them accommodate a push fit for the 1¼-inch (32-mm) ID aluminum tubes used for crossarms, each of which is 4 feet (1.2 meters) long.

To avoid some of the weaker points in antenna construction much care must be used in marking the boom for drilling. Nothing but utmost pains is good enough. Circumscribe on the outside of the metal boom a circle that *exactly* defines the outside contour of the tube hole. If you are using 1¼-inch (32-mm) ID tubing, the circles will be 1.376 inches (35.0 mm) diameter (for a wall thickness of 0.063 inch, or 1.5 mm). A tube with a thicker wall will need holes slightly farther apart.

When correctly marked, the holes will be drilled to a diameter of 1¼ inches (32 mm). When using a large drill through a thin-wall tube, the inside of the holes will be rough and uneven. It is essential that the holes make a perfectly smooth fit for the tubing forming the antenna cross arms.

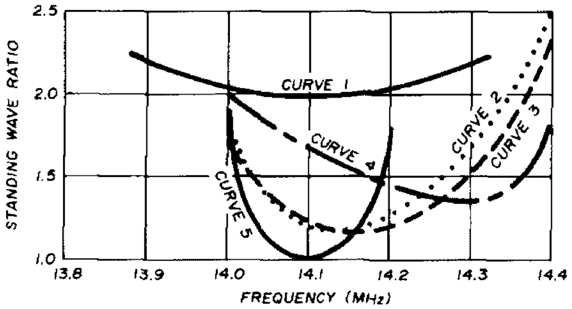
The hole now lies about 1/32 inch (1 mm) inside the circumscribed circle. A steel plug 1.413 inches (36 mm) in diameter is now prepared. File the rough-drilled hole carefully until it is perfectly circular, so that the steel plug fits the hole exactly.

measurement notes

You'll notice that dimensions for holes and parts are given quite exactly, and this is important. If the antenna is to withstand the weather, care in making measurements is imperative. Much care and patience at this point will pay off in a maintenance-free installation. It's surprising how exactly the fitting process can be done, especially if a jig or template is available.

The remaining 1/32 inch (1 mm) of the crossarm hole is now filed out, using the circumscribed circle on the metal and a piece of the crossarm tube as templates. A really tight fit is the objective here – a

joint that will not work loose. The corresponding hole on the other side of the boom is similarly marked and drilled (don't try to drill straight through the boom). By careful measurement, the holes will be so closely opposed to each other that the spreader tips, when assembled, will not be more than 1 inch (25 mm) out of alignment.



Curve 1	Driven element only, 15 feet (4.6 m)
Curve 2	1st director added, 15 feet (4.6 m)
Curve 3	2nd director added, 15 feet (4.6 m)
Curve 4	Boom lifted to 30 feet (9.1 m)
Curve 5	Reflector added, 75 feet (22.9 m)

fig. 2. Standing wave ratio readings taken during assembly of the array. Bridge was located at the junction of the coaxial transmission line and the element except for curve 5, when the bridge was between the coaxial transmission line and the transmitter. Dimensions refer to boom heights above ground.

machining the boom

The only part of the work that required commercial help was turning a 24-inch (61-cm) length of aluminum tubing to join two 15-foot (4.6-meter) lengths of aluminum irrigation pipe used for the boom. The pipe arrived in two 20-foot (6.1-meter) lengths. However, the boom went together so easily, and was erected so simply, that if I did the job again, I'd have no hesitation at all in using a 40-foot (12.2-meter) boom.

If, as in my case, the pipe is not absolutely circular, carefully measure the maximum and minimum diameters, average them, and subtract 0.003 inch (0.1 mm) for clearance. Grease the pipe well (a good silicone grease is best), and hammer the two boom lengths onto the junction piece using a block of wood and a sledge hammer. This completes the boom except for minor details.

spreaders

Fiberglass tapered tubes are used for the spreaders, rather than aluminum, because the fiberglass causes less wind loading. The spreaders are 13 feet (3.96 meters) long, 1¼ inches (32 mm) in diameter, tapering to 3/8 inch (10 mm). This configuration presents a nicely balanced combination of adequate stiffness with light weight.

At a length of 13 feet (3.96 meters), the two spreaders in line need no separation to achieve the desired length, and so are simply moved down the 1¼-inch (32-mm) diameter crossarms until they meet at the center. They are fixed in this position by the bolts that secure each crossarm to the square plate, since these bolts pass through the aluminum tube, fiberglass tube and plate.

The only other adjustment is at the point where the spreader emerges from the crossarm. Since the spreader is tapered, some play will exist at this point. This play is taken up by inserting a neoprene O-ring just inside the end of the aluminum crossarm. It needs no holding in place, but the joint at this point can be covered by a machined aluminum sleeve, made to fit the two tubes, and secured by two self-tapping screws.

The size of the thick end of the spreader should be checked with the manufacturer before buying your crossarm material. Mine required a little work with an emery cloth to make a perfect fit.

clips and fasteners

The homemade clip shown in **fig. 1** resulted from many experiments. The clamp is a commercially obtained item. The tube portion, which carries the wire element, is made from 1/8-inch (3-mm) ID copper tubing. The tube radius should be 3 inches (76 mm). Sections of the copper tube are cut off so that the ends point at right angles to each other. The supporting band, to which the 3/8-inch (10-mm) diameter clamp attaches, is a

strip of copper cut to length, formed, and soldered to the copper tube as shown. The idea is to provide support to the wire elements through the entire radius of the bend, presenting no solid point where the wire could fracture from vibration.

When the wire loops have been fed through the tubing clamps, cut to length, and secured, the wire will be prevented from slipping through the clamps by solder dropped onto the wire just clear of the clamp ends. One or two turns of number-22 AWG (0.6-mm) wire at the desired point will assist in making a good solder joint. Note that cold solder jobs in this area are a definite "no-no" if your antenna is to withstand the elements.

Be certain the wire elements move freely within the copper tubing. Any sharp bends in the wire will defeat the entire idea of this project; remember, care in construction will pay off in maintenance-free operation.

rotator

The usual pressure-grip clamp arrangements used on the antenna rotator shaft (2 inches, or 51 mm) diameter are inadequate for an antenna of this size. If a HAM-M rotator is used, the ¼-inch (6-mm) bolt that screws into the rotating shaft at top and bottom of the motor should be removed. This bolt should be replaced with a 3/8-inch (10-mm) high-tensile-strength bolt to avoid shearing under the stress involved.

performance

The curves in **fig. 2** show the standing wave ratio of the antenna during assembly. Curve 5 shows swr with the complete array elevated to 75 feet (22.9 meters). Antenna bandwidth is 200 kHz with an swr of less than 2:1, which is to be expected. Qualitative tests indicated a front-to-back ratio of 28 dB. Not bad for a home-built antenna.

reference

1. Lee Bergren, WØAIW, "The Multielement Quad," *QST*, May, 1963, page 11.

ham radio

automatic reset timer

Using the versatile NE555 timer IC as a ten-minute ID timer or repeater time-out timer

Harvey Vordenbaum, W5ZHV, 5446 South Shady Creek, Houston, Texas 77017

Recently, when a friend expressed a need for a timer with automatic reset or triggering capability, the NE555V integrated circuit came to mind. The NE555V is a versatile, inexpensive IC designed especially for timer and oscillator use. External components determine the time delay or oscillator frequency and duty cycle. The output is capable of sourcing or sinking 200 mA, adequate for most small relays.† Other basic timer circuits have been described previously,[1,2] but all require triggering by a manually-operated switch or by relay contacts.

†Although a relay coil is shown as the load, the NE555V may be able to operate some loads directly. The 200-mA load current is for a supply voltage of 15 volts, however, and load current will be less for lower supply voltages.

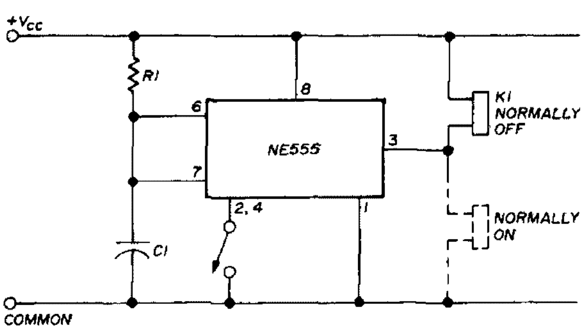


fig. 1. Basic timer circuit uses NE555 IC. Time delay is set by RC product of R1 and C1.

Fig. 1 shows the basic timer circuit. As shown, the relay or other load is normally *off*. For normally *on* operation connect the relay or other load between pin 3 and ground. The delay time constant is set by the simple RC product of R1 and C1. Pins 2 and 4 are grounded momentarily through switch S1, providing start or manual reset of the timer. Pin 2 is the trigger input; when it is grounded momentarily the timing interval is started. Once started it cannot be retriggered. Pin 4 is the reset input; when it is grounded momentarily the timing interval is ended (the output goes low). With pins 2 and 4 connected together both functions are obtained with one push of the switch. When the reset function is not wanted pin 4 should be connected to pin 8, V_{cc}.

fig. 2. Two or more timer ICs may be cascaded when sequential timing is required.*

The NE555V IC can also be triggered by a negative-going pulse applied to pin 2. Thus, two or more timers can be cascaded for sequential timing as shown in **fig. 2.**

*For those circuits requiring two NE555V timer ICs, you might consider using the new dual-timer IC, the NE556, which contains two NE555s in one package. **Editor**

When pin 3 goes low at the end of a timing interval, a negative pulse is generated by the .001-μF capacitor and 27k resistor. This pulse triggers the start of the second timer IC. If a similar pulse circuit is connected to the second timer, and *its output is connected to the trigger* input of the first timer, the second timer automatically triggers the first timer as shown in fig. 3.

The second timer can be set to determine the *on* time of the first timer. When cause the first timer cannot be reset or triggered by the trigger pulse from the second timer while the first timer is in its timing interval.

For time delays of more than a few minutes a good quality tantalum capacitor *such as a Sprague 150D should be* used. For example, a ten-minute time delay requires a 100-μF capacitor and a 6-megohm resistor. Two possible applications for this circuit are station ID timers and repeater time-out timers.

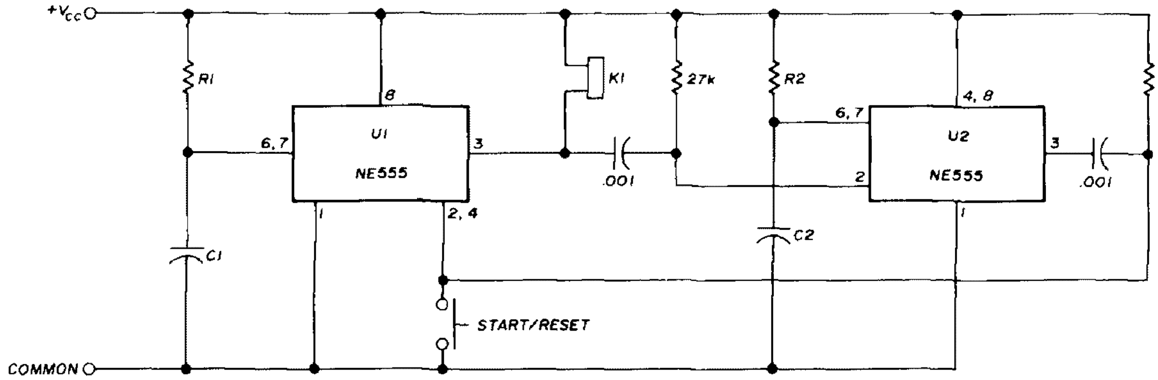


fig. 3. In this circuit the second timer, U2, automatically triggers the first, U1.

the timing interval of the second NE555V is completed a negative pulse is generated from the second .001 μF capacitor and 27k resistor. This pulse triggers the first timer which turns off the relay. Thus, the first timer determines the delay time interval and the second timer determines the *on time* of the relay.

At the end of the *on* time the first timer is automatically triggered and starts another timing interval. If the start/reset switch is closed momentarily, both timers are triggered. This does not matter be-

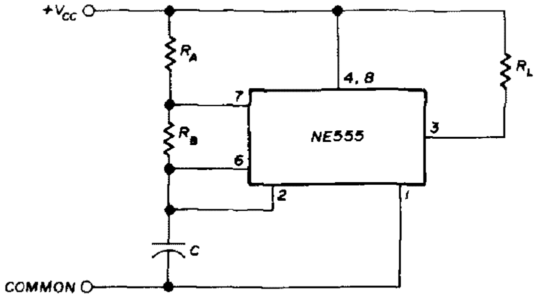


fig. 4. The NE555 may also be connected as an astable multivibrator but this circuit has several disadvantages when used as a timer (see text).

You might also consider using the astable multivibrator circuit (**fig. 4**). The charge and discharge times of the capacitor, C, are determined as follows: Discharge time (when the output, pin 3, is low) = $t2 = 0.685(R_b)C$. The charge time (when the output is high) = $t1 = 0.685 (R_a+R_b)C$. This circuit has two main disadvantages for longer time delays. For one thing, the first timing interval is longer than subsequent ones. This is due to the fact that the capacitor's charge starts from zero for the first time interval but thereafter operates between 1/3 and 2/3 of V_{cc}. The second disadvantage is that there is no manual reset capability. Even if the reset pin is connected to ground through a switch the capacitor will not be completely discharged, due to the time constant of R_b and C.

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1. D. Blakeslee, W1KLK, "Time — IC Controlled," *QST*, June, 1972, page 36.
2. E. Mooring, W3CIX, "Simple Timer," *ham radio*, March, 1973, page 58.

ham radio

enhancing cw reception through a simulated-stereo technique

A simple method for improving copyability of CW signals

Max Blumer, WA1MKP, Box 446, Woods Hole, Massachusetts 02543

A sharp audio filter is a great help in copying CW signals through noise and interference. A filter passband of 100 Hz will accommodate the keying bandwidth at 20 wpm and tolerate some short-term drift of transmitter and receiver.[1] However, such a narrow filter makes scanning of the band slow and difficult, and with high skirt selectivity the character of the received signal and its attack and decay may become distorted from ringing.

threshold gating

Higher noise reduction than possible from a tuned filter can be accomplished through threshold-gating, a technique in which the received and filtered CW signal is used to key a relay[2] or an electronic switch.[3] In turn, the relay or switch feeds the received and filtered signal or a sidetone oscillator to the headphones or speaker. Threshold-gating achieves its selectivity through the switching process. No signal is heard off frequency or between the dots and dashes. However, the original signal is highly distorted, especially in its attack and decay, or is completely replaced by a sidetone. As a result, feel for band conditions and for the quality and "signature" of the signal are lost, and any reply slightly off the filter frequency will not be heard at all.

This discussion suggests that you cannot, at the same time, use high filter selectivity and retain an essentially unaltered CW signal and full feel for the signal and the band. Therefore, sharp filters seem to have little value in contests and net operations, where you want to respond rapidly, often to signals which

appear to one side of your receiver center frequency.

another approach

This is certainly true as long as you listen to the filtered signal only. However, if you were to listen simultaneously to one speaker fed with the processed signal and another speaker reproducing the "raw" CW signal, you could retain a feel for the band and would even hear chirp and clicks that extend beyond the filter

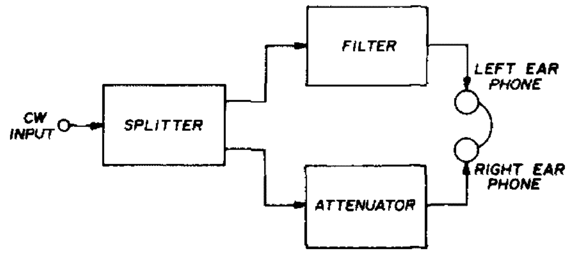

fig. 1. Block diagram showing how a simulated-stereo signal is derived from unfiltered receiver audio containing a desired signal plus QRM and QRN.

passband. Also, if the received signal were drifting you would not lose it, and you could hear replies at some distance from the center frequency of your filter. Obviously, you would also lose some of the advantage of the filter and QRM and QRN would have returned to some degree.

To this end, you *can* be helped by the ability of the brain to discriminate between signals which appear at both ears or only at one. To do this, the incoming CW signal is divided into two channels as shown in **fig. 1**. Half of the signal goes through a sharp filter to one ear, and the other half is passed unfiltered to the other ear. The level or balance control in the unfiltered side compensates for any attenuation of the filter. It is set so that a tone centered in the filter passband is heard by both ears at the same loudness level. This tone is then perceived stereophonically and appears centered within your head.* Any other tone is attenuated by the filter and will be heard predominately from the unfiltered channel. Consequently, all signals that are not passed by the filter appear to come from that side which is receiving the *unfiltered* channel.

*Because this effect may be enhanced or diminished by the phase relationship of the audio in the two earpieces, it would be desirable to try transposing the leads to one of the earpieces to see what happens. editor

results

The effect of this simulated stereo reception of CW signals is dramatic. Interfering signals and broadband noise appear to come from a point off to one side of the head. The desired signal, centered in the filter passband, appears within or just in front of the head and assumes a transparent clarity that is hard to describe. The signal-to-noise ratio is much improved. The character of the signal is preserved and ringing is either absent or less apparent than in monaural reception with the same filter. Chirp and clicks are readily noticed, and even drifting signals or DX signals with multipath distortion are readily copied.

The mind seems to concentrate automatically on the desired signal and to be relatively unaware of and undisturbed by the signals outside the filter passband. Yet, that information is present and an off-frequency reply is heard just as well as if no filter were in use.

practical considerations

I have used this approach with a simple passive toroid filter and with a more complex filter and threshold-gate combination. Both are effective. The latter has an advantage on some occa-

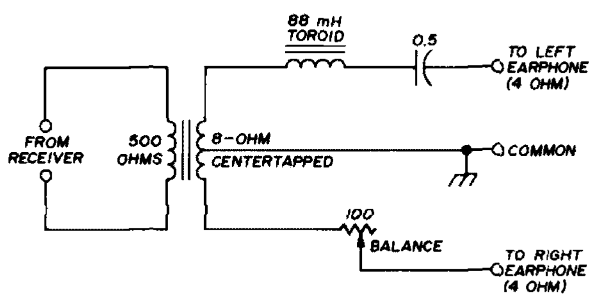

fig. 2. Circuit of complete simulated-stereo setup including simple but effective series-tuned toroidal filter.

sions, but for its simplicity and ease of adjustment the toroid filter is superior. The high-impedance audio signal from my HW101 is matched to the low impedance of the series-tuned toroid filter by a transformer, as shown in **fig. 2**. The center-tapped secondary provides two equal signals, one attenuated by the balance control and then fed to the right channel of a low-impedance stereo headset.

The toroid filter in the other half of the secondary resonates at 790 Hz and has a 3-dB bandwidth of 60 Hz; this frequency was chosen because the apparent stereo separation increases at lower frequencies. The output of the filter is fed directly to the left channel of the headset. The balance control should be adjusted on a moderately weak CW signal; a strong, steady tone, such as from a crystal calibrator, gives a slightly different balance.

First, peak the signal in the filter by listening only to the left earphone, then put on both phones and adjust the balance control until the signal appears centered. The range of the balance control is sufficient to move the signal from the far right across center to the left. Little further adjustment is required under differing band conditions.

The principle can be applied in various ways. Other input and output impedances can be accommodated with different transformers, or a parallel-tuned toroid filter (approximately 500 ohms) could be used. Other filters could be substituted, and instead of earphones a stereo amplifier and speaker combination used to give a good demonstration of simulated-stereo CW reception to interested listeners at a club meeting or hamfest.

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2. R.M. Myers, W1FBY, "Recent Equipment: The Douglas Randall Scrubber," *QST*, November, 1971, page 51.
3. J.J. Duda, W2ELV, "Noise Reduction for CW Reception," *ham radio*, September, 1973, page 52.

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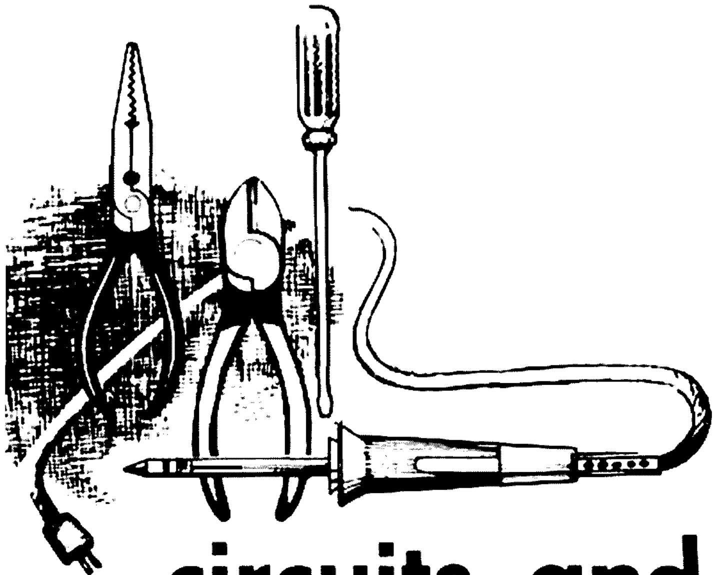


circuits and techniques

ed noll, W3FQJ

storage-battery QRP power

In a self-sufficient QRP station the storage battery is king. Although battery power need not be limited to low-power stations, QRP operation does permit you to gain knowledge of the science, gradually weaning you away from plug-in power, and over a period of time you can make your amateur station reasonably independent of the power mains. Solar power, wind power, etc., can make it so. The wind and solar conversion combination is attractive because on very cloudy days when there is no bright sunshine, there often is wind. Initially it might be expedient to take some power from the mains with an efficient battery charger, particularly when power demand is above the QRP level and there is a sequence of dark, windless days.

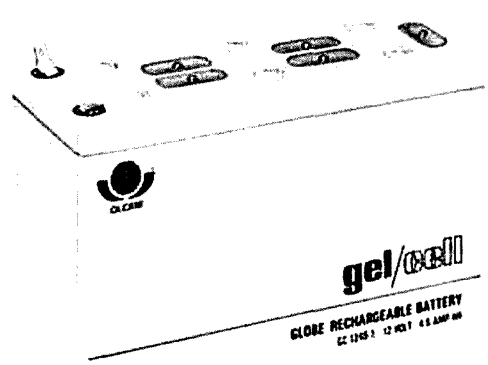
The radio amateur may be able to make a considerable contribution by demonstrating how energy can be derived from light, wind and other means. A house that is partly or completely self-sufficient in terms of heat and electrical power is a worthy objective.

Battery power at kilowatt and higher levels requires a well-ventilated hut or battery room, if lead-acid wet batteries are to be used. During the charge cycle batteries release some hydrogen gas which, if permitted to accumulate, becomes explosive. More gas is released when charging at a high rate or when you permit batteries to overcharge. Normally there is no hazard if the hydrogen gas circulates and intermixes freely with the atmosphere.

A ventilated window box will do for somewhat lower power levels. However, batteries must be selected to withstand the weather extremes of the site or mounting position. Nickel-cadmium batteries have fewer such problems and can be operated in more confined areas but are considerably more expensive at the higher power levels as compared to lead-acid types.

Some of you who lived in farm country before rural electrification may recall the small battery room just away from or attached to farm houses. Remember the wind chargers, small gasoline engines and dc generators? This is not so much nostalgia as some very practical dreaming as to how to obtain at least some degree of self-sufficiency, avoiding some of our

fig. 1. This gelled-electrolyte, 12-volt, 4.5 amp-hour battery is a relative newcomer to the rechargeable battery scene (photo courtesy Globe-Union).



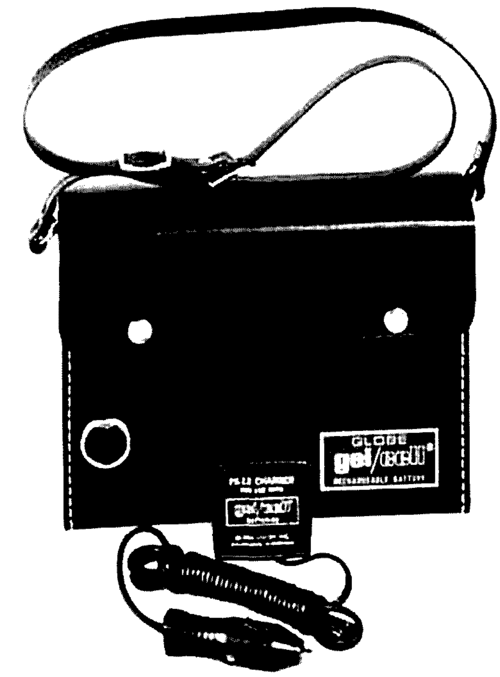


fig. 2. Case and charger for gelled-electrolyte batteries (photo courtesy Globe-Union).

enslavement to mass energy sources plus the high cost that shortages trigger.

small lead-acid batteries

In the realm of QRP operation there are a variety of small lead-acid types. Visit your local motorcycle shop or take a look at the variety of types listed on the motorcycle batteries list in the Sears catalog. Two- and four-ampere-hour (Ah) types can be purchased at low cost. A typical price for a 4-Ah, 12-volt battery is not much more than 10 dollars. Such a low-powered battery displaces very little hydrogen and can be brought into your radio shack with little hazard. It can be charged by the smallest of chargers or by a small solar energy converter. On the basis of a 20-hour discharge period, the 4-Ah battery can supply 200 mA continuously for a period of 20 hours (4/20. Based on amateur operating practice it would be no problem to supply up to 10 watts for almost 8 hours of continuous operation without requiring a recharge.

Power capacity and current capability can be increased by connecting two or more of these small batteries in parallel. Much depends on your operating practices. In most cases operating time is substantially less than the projected maximums suggested by the previous figures.

gelled-electrolyte batteries

There is an attractive newcomer on the scene. It is a lead-acid battery that uses a gelled electrolyte, fig. 1. It is truly a portable lead-acid battery that can be mounted at any angle and, in some models, even charged at any angle. Others charge more efficiently with the battery upright but can be charged at other angles with some limited decline in the total number of recycles. The electrolyte is unspillable and lasts for the full life of the battery, avoiding maintenance and handling problems. The battery has a one-way vent that serves as a release when there is undue gas pressure, although in this style of battery there is much less gassing.

The gelled-electrolyte battery handles

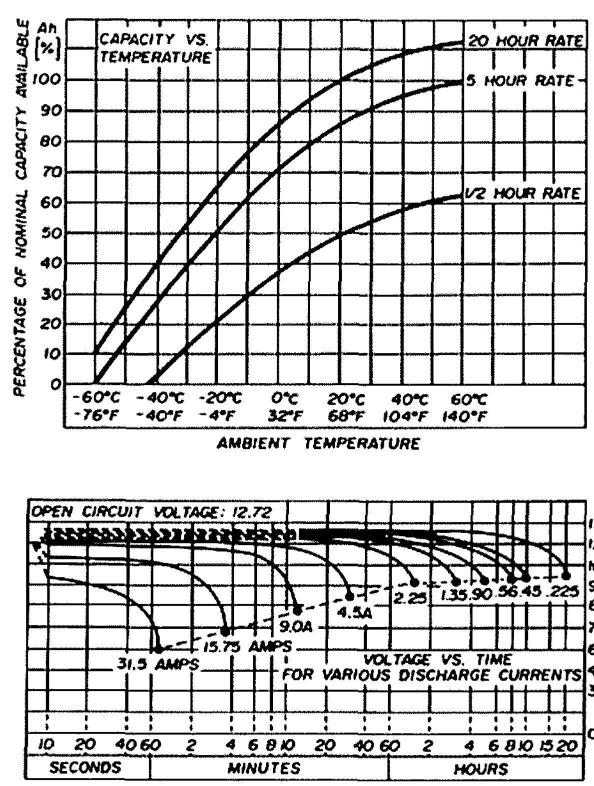


fig. 3. Operating characteristics of a 12-volt gelled-electrolyte battery, the Globe GC 1245-1.

temperature extremes very well and is capable of performing down to -76°F (-17°C). It is tolerant of both overcharge and a deep discharge and provides long, maintenance-free shelf life. Batteries may be connected in series, parallel or series-parallel combinations to obtain desired voltage and current capability. Case and charger are provided for some types as shown in fig. 2.

The characteristics of the Globe-Union 4.5 ampere-hour, 12-volt battery are given in fig. 3. This battery would be a good choice for up to 5-watts QRP operation. Typical capacity figures are shown in item 2 of the specifications chart (fig. 3). Note, under item 9, the charge retention ability of the battery. When sitting unused for six months the charge drop-off is only 82%.

The first graph shows battery capacity as a function of the discharge rate. When discharged at the 20-hour rate the battery provides 100% capacity if operated at normal room temperature (about 69°F or 21°C). The percentage is lower for faster discharge rates.

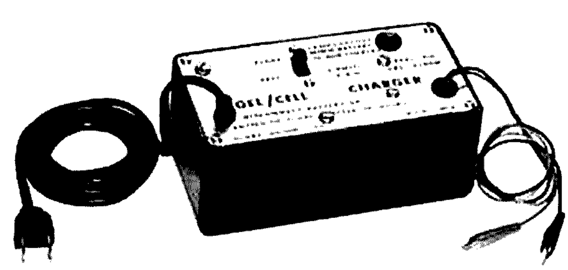
Curves for various discharge rates are shown in the second graph. The top curve, representing the 20-hour rate, indicates a voltage decline to about 10 volts when the current demand is a continuous 0.225 amperes for a period of 20 hours. The final power delivered under this situation is about 2.25 watts (0.225 x 10). If a current demand of 1 ampere is made, note that the battery will provide almost four hours of continuous operation. This is about 10 watts of power for a continuous 4 hours. Thus, the 5-watt rating is a very conservative one.

In normal amateur operations you would have no trouble supplying 10 watts input to a QRP transmitter, and perhaps even more if you take care of the battery, preventing it from overcharging or dis-

GC 1245-1 Specifications

Nominal voltage	12 volts (6 cells in series)
Nominal capacity at:	
225 mA (20-hour rate) to 10.5 volts	4.5 Ah
430 mA (10-hour rate) to 10.26 volts	4.3 Ah
780 mA (5-hour rate) to 10.14 volts	3.9 Ah
2400 mA (1-hour rate) to 9.6 volts	2.4 Ah
Energy density (20-hour rate)	0.96 watt-hours/cubic inch
Specific energy (20-hour rate)	12 watt-hours/pound
Internal resistance of charged battery	approximately 60 milliohms
Maximum discharge current with standard terminals	80 amperes
Operating temperature range:	
Discharge	-76°F to +140°F
Charge	-4°F to +122°F
Charge retention (shelf life) at 68°F	
1 month	97%
3 months	91%
6 months	82%

fig. 4. This charger for gelled-electrolyte batteries provides either fast or float charging (photo courtesy Globe-Union).



charging to too deep a level. You can protect the battery's welfare by keeping an eye on its output voltage under load.

Power demand can be stretched even further if the battery is kept on a continuous floating charge with the permanent connection of a small charger, **fig. 4**. Current demand in amperes can be made for short periods of time.

The above operation also applies for daytime operation when using a solar energy converter as a float charger. Battery charge must be restored before the end of the day if capacity for night-time operation at lower power level is required.

The rated capacities (20-hour basis) for various Globe-Union gelled-electrolyte batteries are shown in **table 1**.

Prices are higher than for comparable wet electrolyte lead-acid types, but substantially lower than the cost of nickel-cadmium batteries. The combination shown in **fig. 2** is especially attractive for use with portable transceivers because it includes a battery-case and a charger. Batteries of this type do vent some hydrogen at the end of the charge cycle, or upon overcharge, and although they are less hazardous than wet electrolyte types, sensible ventilation and avoidance of sparks are advisable during the charging interval.

solar power as a charger

The solar energy converter has excellent battery charging capabilities. There are a number of ways in which you can

table 1. Rated capacity of various Globe-Union gelled-electrolyte batteries.

part number	nominal capacity
GC 210	0.9 Ah
GC 410	0.9 Ah
GC 610	0.9 Ah
GC 1215	1.5 Ah
GC 620	1.8 Ah
GC 426	2.6 Ah
GC 626	2.6 Ah
GC 1245	4.5 Ah
GC 660	6.0 Ah
GC 280	7.5 Ah
GC 680	7.5 Ah
GC 12200	20.0 Ah

use such a source. If the converter has adequate capacity it can even handle the initial charging of the battery. This is really not too much of a requirement if you are willing to charge the battery initially in a series of long-term steps.

The second technique is to use a conventional charger to obtain the initial

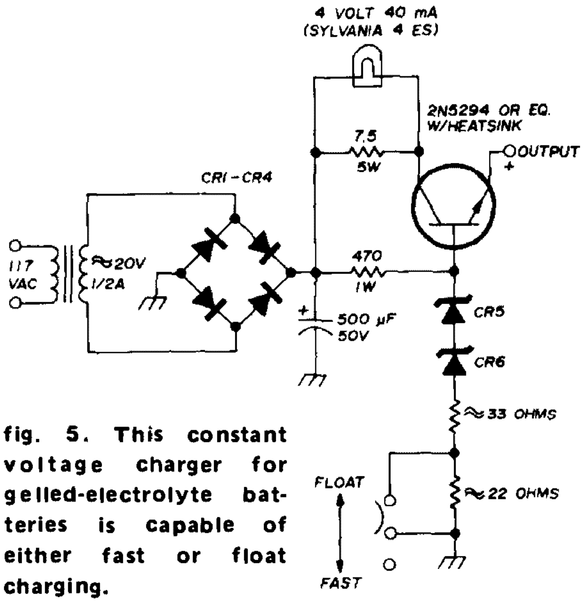


fig. 5. This constant voltage charger for gelled-electrolyte batteries is capable of either fast or float charging.

charge. Then the solar energy converter can be pressed into service as either a float or trickle charge source. In this mode of operation the current demand is only one-quarter or even a smaller fraction of the initial charge current requirement. If the battery is critical of charge voltage or current it is possible to add a solid-state regulator to the output of the solar energy converter. Under less stringent requirements the protective diode that is a part of the solar device can prevent the discharge of the battery when the impinging light on the solar cells is inadequate to maintain the proper charge current.

battery chargers

There are several battery-charging arrangements relating to battery types and mode of operation. Insofar as the initial charge is concerned, some batteries can be charged quickly while others are preferably charged at lower current rates over a longer period of time. In general,

nickel-cadmium types are charged at a slower rate than lead-acid types. Gelled-electrolyte types are usually charged at a slower rate than wet electrolyte cells.

For practically all types of rechargeable batteries the slow charge is much preferred over the fast charge, although certain battery types are less ill-affected by a fast charge, and in some circumstances you may have to sacrifice some battery life in favor of fast charging. Regular amateur radio operations are such that you can usually take advantage of slow charging, and, therefore, gain an extension in battery life.

table 2. Charging-current values for Globe-Union gel-cells.

battery rating	maximum initial charge current	approximate final current
0.9 Ah	0.15 Amp	10-20 mA
1.5 Ah	0.25 Amp	20-40 mA
1.8 Ah	0.30 Amp	20-40 mA
2.6 Ah	0.40 Amp	30-60 mA
4.5 Ah	0.70 Amp	50-100 mA
6.0 Ah	0.90 Amp	60-120 mA
7.5 Ah	1.20 Amp	80-160 mA
20.0 Ah	4.00 Amp	100-300 mA

Batteries can be charged and then discharged to a specified end voltage. At this time the battery is again charged fully, discharged, etc. In this mode of operation the battery is on charge whenever it is not being discharged by a connected load. Two other arrangements keep the battery on continuous charge. These are known as a constant-voltage charger (float voltage charge) or a constant-current charger (trickle charge). In the float voltage system preferred for gelled-electrolyte batteries the charge voltage is held constant while the current is free to vary. In contrast, the trickle charge plan preferred for nickel-cadmium batteries maintains a constant charging current while the voltage is allowed to vary.

The chart of table 2 shows the initial charge current and fully charged current for the standard ratings of various Globe-Union gelled-electrolyte batteries. For example, the 4.5-Ah, 12-volt battery begins charging at a level of 700 mA. Full charge is indicated when the battery voltage reading is 14.4 volts and the charge current has dropped to a level between 50 and 100 mA. This corresponds to a final cell voltage of 2.4 volts.

When the same gelled electrolyte batteries are to be kept on continuous charge it is preferable that the charge voltage be held to 2.25 volts per cell, or for the 4.5-Ah version, a final voltage of 13.5 volts. Therefore, the charger must supply a constant 2.25 volts/cell (13.5 volts in the case of the 4.5-Ah, 12-volt battery).

To obtain the maximum number of recharge cycles the on-charge voltage initially should be such that the battery charge is brought up to 2.4 volts per cell. This charge should be continued until the current drops to the values shown in the tables. At this point the charger should be switched over to a float voltage of 2.25 volts per cell.

In practice as many as 200 to 400 full charge/discharge cycles are possible. If a float voltage charge is maintained, instead of permitting complete discharge, thousands of cycles of operation are feasible.

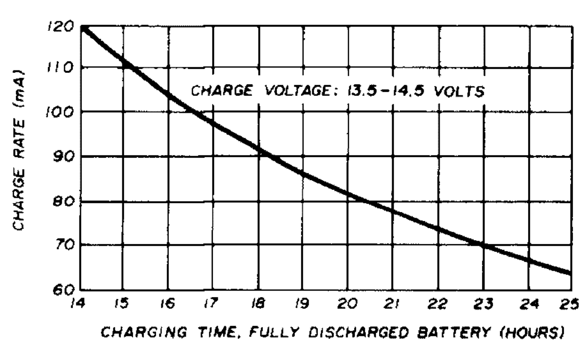


fig. 6. Battery charge data for the Eveready N86 nicad battery.

The charger shown in fig. 4 permits both modes of charging. Note the switch that can be used to select either float or fast charge. An indicator lamp is turned on when the battery reaches 80% of full charge.

An example of a constant-voltage

charger that can be used for float or fast charge activity is shown in fig. 5. For fast-charge activity it provides exactly 14.4 volts. The constant voltage is maintained by the series power transistor and two series-connected voltage-regulator zener diodes. The precise value of the constant voltage is set by the two resistors connected between the anode of zener diode CR6 and common. In the float position only one resistor is used, the output voltage being maintained at 13.8 volts. An additional resistor is inserted in the circuit when the fast charge voltage of 14.4 volts is desired.

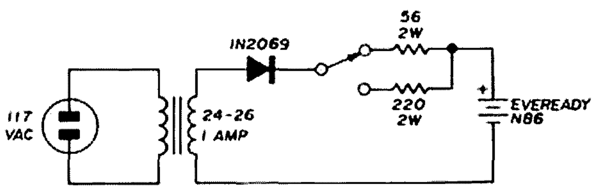
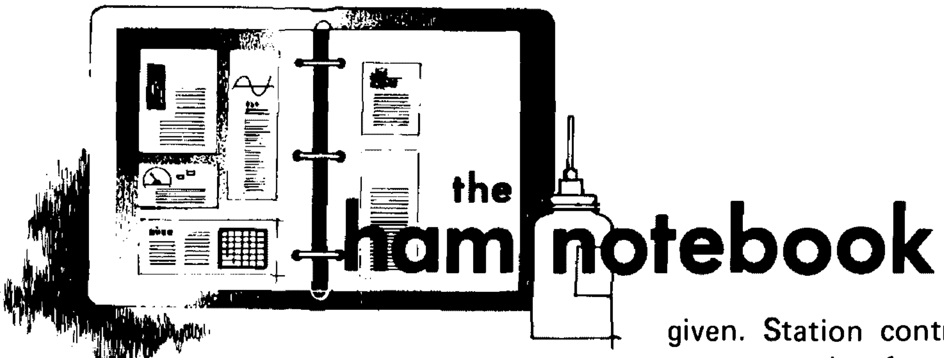


fig. 7. Basic constant-current battery charger for nicad batteries.

A nickel-cadmium battery is best charged with a constant-current source. A rather extended charge time is recommended and it is advisable that the charge rate not exceed the 10-hour figure. Information for the Eveready N86 12-volt battery (refer to information given in the August column) is shown in fig. 6. A series charge current of 120 mA will charge the battery fully in 14 hours. About 82 mA of charging current will do the same in 20 hours.

A basic charging circuit for nickel-cadmium batteries is shown in fig. 7. The circuit is fundamental although the values given are for the Eveready N86 battery. This circuit provides a charging current of 120 mA. After a battery has been fully charged, a trickle charge arrangement can be used to maintain full charge. Recommended trickle charge current for this battery falls between 24 and 40 milliamperes. This current value can be obtained by switching a higher value resistor into the constant-current charging current.

ham radio



transmitter fail-safe timer

This timer was designed to provide for the disabling of an automatic RTTY transmitter in the event of control failure. In particular, this station has features which allow control codes to be punched on tape. After starting a tape the operator may leave the room to work on one of the many unfinished projects always waiting in the basement. Another feature allows a remote operator to punch tape on the LRXB composite tape set and then have it automatically replayed at its completion (operator on premises, but not necessarily at the control console). In either case a tangled tape or a mispunched control code could leave the transmitter keyed. There is also the possibility of the operator leaving the switching set-up in such a way that the system can not function properly. In all such installations it is necessary to provide some sort of backup control so that the channel is not blocked by a stalled automatic system.

This system allows the transmitter to remain keyed no longer than ten minutes (five minutes for the replay). If the time limit is exceeded, the transmitter is removed from the air and cannot be rekeyed until reset by the control operator. The timer is reset by the CW identification device. Hence, it is possible to make transmissions longer than ten minutes, but only if the proper identification is given. Station control wiring prevents a remote station from inserting the CW ID sequence on a replay tape. However, such codes may be inserted when the control operator punches tape.

Originally I planned to use a NE555 timer IC to set the ten-minute time period. However, I found it impossible to reach the ten-minute limit with the capacitors I had on hand. Since such capacitors were also expensive and somewhat difficult to find, it was decided to go to a shorter period and use a 7490 divider to extend the period. This approach costs no more than the original plan with a high quality/cost capacitor, and it allows for other timing periods as well.

The NE555 oscillator is set at approximately one cycle every 1.4 minutes. The output is introduced to the 7490 decade divider. Pin 11 of the 7490 goes high on the eighth count (approximately 11.2 minutes) after the 7490 is enabled by U4B. Then the output of U3D goes low, forcing the output of U3C high. When pin 8 of U3C goes high, Q1 conducts, removing the base bias from Q2. When Q2 stops conducting, the disable relay opens, resetting a holding relay in the control unit and disabling the entire automatic system. The Q1, Q2 configuration was chosen so that the disable relay would be normally on and power supply failure would also reset the system.

When the replay is enabled U3A, pin 3, goes high. Pin 8 of U2 goes high on a count of four (approximately 5.6 minutes); U3B goes low, shutting down the disable relay as previously described.

The 7490 is enabled when the transmitter is on (U4D pins 12 and 13 low). If, however, pin 3 of U4A is forced low (by the CW ID device), then U4B pin 6 goes high, resetting the 7490 to its zero count. It has not been found necessary to reset the oscillator, but this may be done for increased accuracy in the timing periods.

If either the transmitter or the CW ID is keyed, Q3 conducts, keying the *transmit* relay. Also, when the CW ID is keyed, Q4 conducts, keying the CW ID relay which is used to disable the station tape readers and keyboard during the identification sequence.

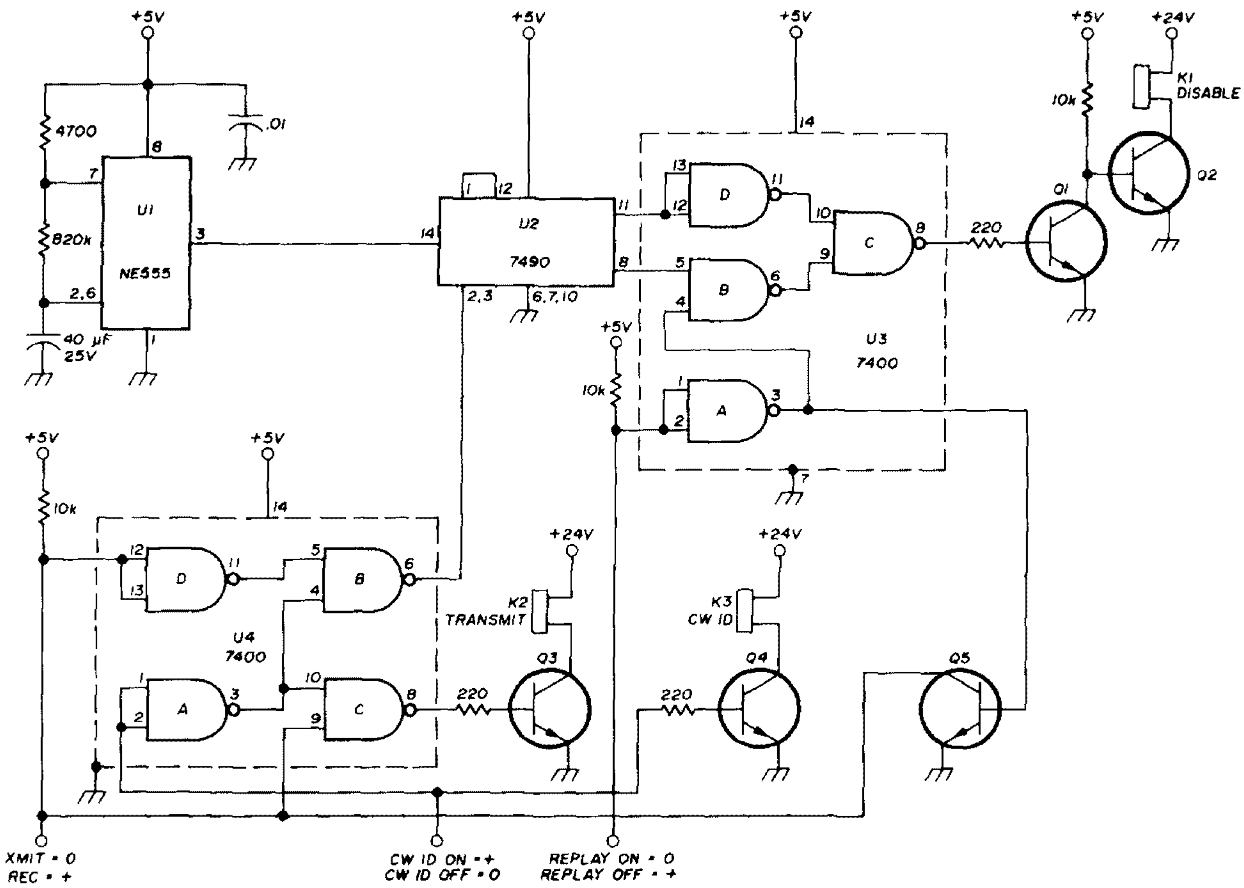


fig. 1. Circuit for the fail-safe timer. Nearly any npn transistors may be used at Q1 - Q5. Relays are Potter and Brumfield KH4703, 24-volt coils.

When the replay is keyed, Q5 pulls the transmitter line low (a single transistor was used here to eliminate the need for another IC package), enabling the 7490 and keying the transmitter. As mentioned above, keying the replay allows the four count from the 7490 to reach Q1 via U3, setting the time-out period for the replay at half the value chosen for operator initiated transmissions.

A revision planned for the future would include an *idle line* detector to sense the lack of regular transitions from mark to space on the loop to time the disable relay out in sixty seconds. This would decrease the recovery time for the channel in the event that a tape reader is not properly enabled or some other malfunction keys the transmitter, but does not allow any data to be sent.

The circuit described here has been in operation for a period of four months. It has never timed out a transmission of proper duration and on several occasions it has removed the transmitter from the air when transmissions were too lengthy (and a few times when the proper control codes were not received by the system).

The logic draws less than 100 mA at five volts. The relays are 24-volt P&B KH4703 and are operated from the un-

regulated dc supply. This supply is 18 volts under the load of the five-volt regulator, and is quite adequate for reliable operation of the relays. The unit is constructed on perf board and is enclosed with the CW ID device in an rf-proof enclosure. All leads entering or leaving the enclosure are bypassed.

Notice that since the timer is reset by the CW ID, the timer would not function properly if the CW ID were inserted automatically every eight to ten minutes. The CW ID must be inserted (possibly on tape) by the control operator for the transmitter to run longer than the limit set by the timer.

Robert Clark, K9HVW

waveguide klystron cooler

In many amateur microwave assemblies, it is impractical or even impossible to obtain adequate air flow around reflex klystrons or other signal sources. One solution is a simple water-cooled section of waveguide. Such a solution is more practical than fabricating a water jacket for the klystron itself because the guide section can be used with any flange-mounted klystron or Gunn diode. An example unit is illustrated in **fig. 2.** It was fabricated from a short section of WR-90 guide with UG-39/U flanges and 1/4-inch (6-mm) water lines. The klystron, a Varian 6975, was operating with a beam power of 9.0 W. The temperature reduction at the worst case was greater than 15°C (see **fig. 3**). The coolant flow rate was 0.6 liters/minute, coolant temperature was 25°C.

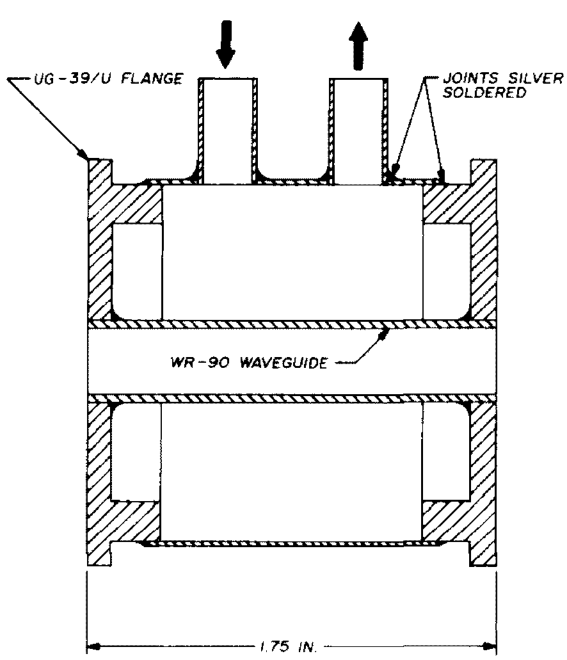


fig. 2. Water-cooling jacket for WR-90 waveguide. Input and output tubes are ¼" (6 mm) diameter.

John M. Franke, WA4WDL

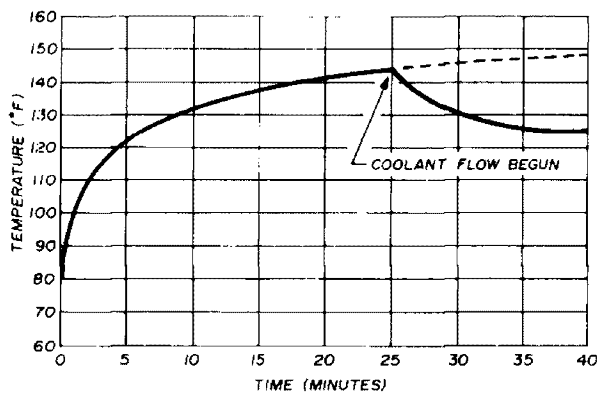


fig. 3. With coolant flow rate of 0.6 liters/minute (25°C coolant temperature) klystron temperature decreased 15°C, minimum.

protector guards against auxiliary battery drain

The latest Sears catalog has an item which should be of interest to amateur operators who use multi-battery electrical systems for mobile, portable or repeater use. The Sears system, which sells for less than $15.00, charges the auxiliary battery, without switching, while the engine (or recharging system) runs. It also protects the engine-starting battery from discharging while the engine is not running such as when operating portable or camping. The system, which consists of a heavy-duty 50-amp circuit breaker and a solid-state switching arrangement, can be used with two or more batteries wired in parallel, 6 to 32 volts, negative ground. The Sears system can be ordered from your local Sears store or sales office; order catalog number 28T7105.

Jim Fisk, W1DTY

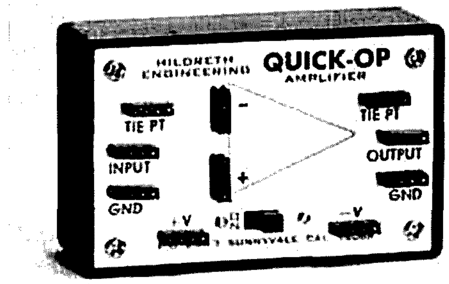
new products

vhf test meter

The new Ascom ASMR100 transmitter/antenna system tester is a three-way test instrument which provides a constant monitor of transmit/antenna functions in any vhf system. When installed in the transmission line between the antenna and transceiver, it provides an accurate check of power output with settings of 25- or 50-watts full scale. Accuracy is ±5 percent. The second in-line function of the instrument measures the standing wave ratio of the antenna circuit to assure optimum efficiency. The instrument can also be used as a field-strength indicator.

Attractively packaged in an all-aluminum case to resist rust or corrosion, the Ascom system tester is designed to mount permanently or for use by servicemen as a precision test unit. Frequency range is 144 to 174 MHz. Suggested resale price is $69.95. For more information write to Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106, or use *check-off* on page 94.

quick-op amplifier



The new Quick-Op Amplifier now available from Hildreth Engineering provides an easy way for you to experiment with a wide range of operational amplifier circuits. Solderless connectors are provided on the front panel of the device so you can easily install resistors, capacitors and diodes, building any op-amp circuit you wish. The internal 741C op amp IC has been tested for gain and dc offset adjustment. Two standard 9-volt transistor-radio batteries mount inside the case, making a completely self-contained unit. If you've been wanting to use op amps, but haven't yet, here's a good way to get started. $11.95 each, less batteries, first-class postage paid, from Hildreth Engineering Company, Post Office Box 3, Sunnyvale, California 94088. For more information, use *check-off* on page 94.

dBm/microvolt conversion scale

A free conversion scale is available from Singer Instrumentation that converts microvolts to dBm and vice versa. High resolution is afforded by the 10-inch (25-cm) scale length. The scale is printed on silver foil with adhesive backing suitable for permanent attachment to any instrument.

Write to Singer Instrumentation, 3211 South LaCienega Boulevard, Los Angeles, California 90016, or use *check-off* on page 94.

More Details? CHECK–OFF Page 94

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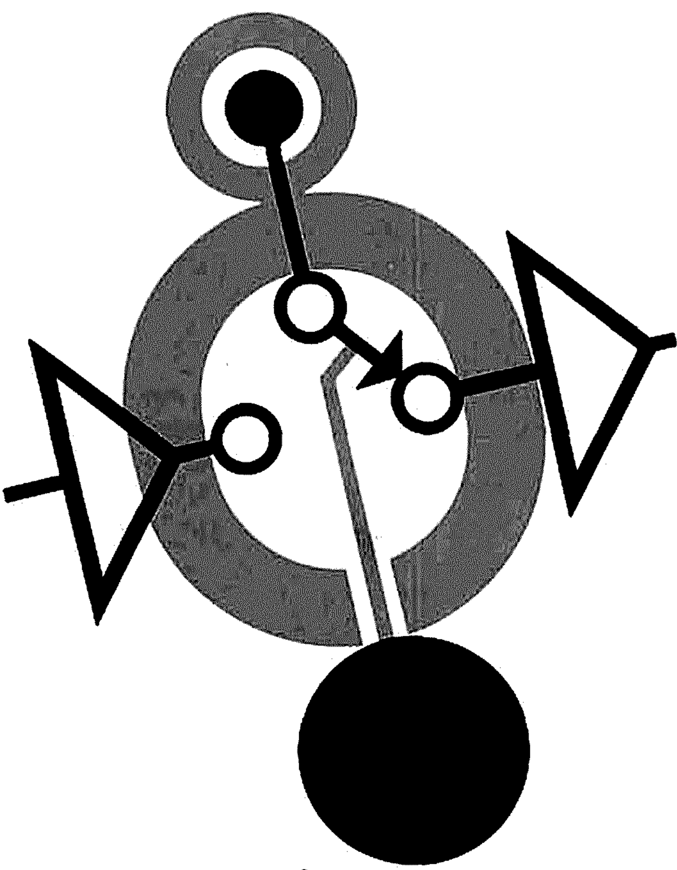
NOVEMBER 1974



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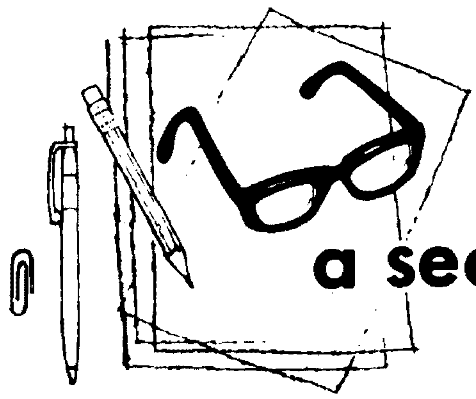


a second look

by jim fisk

Here's a new technology that you're going to be hearing a lot more about in the not too distant future: *additive* printed circuit wiring. If you're wondering what's new, the process used to manufacture most of the PC boards used nowadays is called the *subtractive* process: you start with a copper-clad board and etch out the desired conductor pattern. In the additive process the copper is deposited on the board in the desired conductor pattern. The big advantage is that the process uses up to 75% less copper — obviously important to an industry that is facing a real squeeze by the world-wide copper shortage.

Additive printed-circuit wiring has been around for several years, but until recently users have been troubled by peeling and edge lifting, as well as warpage and cracking. Now, however, many of these problems have apparently been solved, and it appears that the additive technique is about to produce a substantial portion of the PC boards now manufactured by the conventional subtractive process. In fact, some industry spokesmen indicate that the additive process will account for up to 50% of the market by the end of 1975.

In addition to the conservation of copper, the additive printed-circuit process eliminates undercutting due to etching and offers improved circuit design capabilities such as closer line-width tolerances. For example, in standard etched printed-circuit boards with 1-ounce copper foil, the minimum line width is about 10 mils; with the additive process 5-mil line widths are routine. Closer manufacturing tolerances in the near future are expected to allow line widths down to 2 mils.

In addition, when used for high production quantities, the cost of boards produced by the additive technique are considerably less expensive. In mass applications such as television sets and stereo systems, the savings over subtractive PC boards can be as much as 35%. And high production is not the only promising area for additive circuits — because of the smaller line widths and spacings the technique is expected to find wide use in high-density interconnection applications such as computer peripherals.

The principal problem that had to be solved before the additive technique became a viable manufacturing process was finding a way to make the electroless-copper stick to the glass-epoxy laminate. This adhesion (typically 8- to 10-pounds pull) is provided by micropores on the laminate surface. This is no easy task, but one successful approach is to coat the bare laminate with a special adhesive which is activated by a chromic-acid etch. The etchant produces micropores in the adhesive surface and is similar to the treatment used in the plating of plastics.

I don't expect that this technique will replace the various subtractive systems that amateurs have developed to make their own printed circuit boards, not in the near future anyway. However, printed-circuit boards were being used by the manufacturers of television sets long before the process was even considered for one-off amateur projects, so nothing will surprise me. Perhaps one of the chemists among us will come up with a simple additive PC process that we can use in our own home workshops.

Jim Fisk, W1DTY
editor-in-chief

presstop

FCC AMATEUR RADIO ADVISORY COMMITTEE PROJECT, proposed in ARRL/FCC meeting May 10th beginning to warm up as Amateur and CB Division staff lay groundwork for a formal presentation to the Commissioners. A letter has been sent to quite a number of amateurs outlining the purposes and responsibilities of the ARAC.
Letter Was Definitely NOT An Invitation to membership in the ARAC, but instead was sent out as a means of determining whether there would be sufficient support from amateurs to insure ARAC success if the Commissioners were to approve it. Letter proposed basic 12-15 member "Steering Committee" to be chaired by FCC representative, a number of Ad Hoc subcommittees (each chaired by a Steering Committee member) to consider specific problem areas. Steering Committee membership would be by invitation of the Commission, with broad geographical as well as interest-area representation.

ARRL PREPARATIONS FOR 1979 World Telecommunications Conference should be aided by W1RU and K1ZND having participated in a frequency management seminar in Geneva during September.

BLOCKBUSTER TRANSCEIVER FROM HEATHKIT announced this month (see centerfold). The new SB-104 is all solid state, totally broadbanded, digital readout and covers all ham bands from 3.5 to 29.7 MHz! Delivering 100 watts SSB or CW, the SB-104 operates from 13 Vdc, weighs only 20 pounds and measures about 6x14x14 inches; its 6-digit readout provides 100-Hz resolution, 200-Hz accuracy.
New Accessories Complement the SB-104 -- SB-230 linear with conduction-cooled output tube; SB-614 Station Monitor; SB-634 Station Console; SB-644 Remote VFO. Truly state-of-the-art!

GENAVE GOING TO DIRECT SALES IN THE AMATEUR MARKET, with change becoming official November 1st. Inflationary pressures, determination that 75-80% of Genave's ham gear has been selling by mail order anyway are cited as the reasons. High proportion of mail-order sales has meant most service has been factory responsibility, so change to factory selling is a logical one.
Genave's Pricing will reflect the change -- list on the popular GTX-200 will go from $399.95 to $299.95, and other models will show similar reductions.

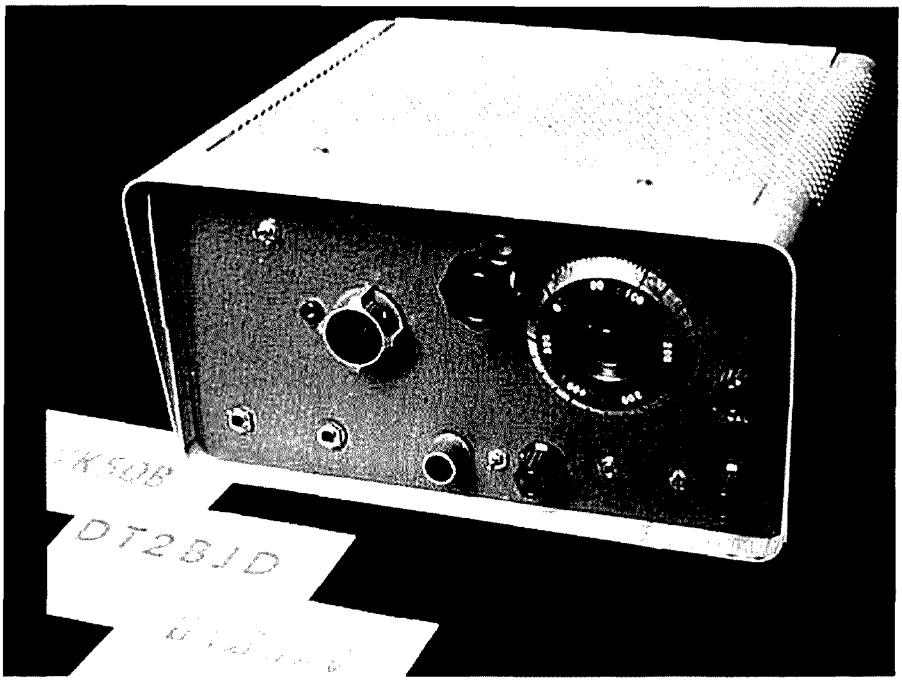
HILDA WETHERBEE LEAVES HAM RADIO. Hilda, a key part of HR since its beginning and well known to many amateurs and the ham industry for her participation in many hamfests and highly effective ad selling, has joined a research firm in New Hampshire.

MOBILERS BEWARE: Minnesota has joined Florida and Virginia in forbidding the wearing of headphones while driving.

ANOTHER 10-METER BEACON PLANNED, this one proposed for San Diego by K6HME and WB6KNC. Request for Special Temporary Authority has been returned for justification of proposed 100-watt power and SSB ID, but as soon as questions are resolved it should be operative using K6HME call. Proposed frequency is 29.0 MHz, same as German DLØAR beacon.

"COMMERCIAL" USE OF HAM BANDS subject of concern in Washington, confusion among hams. Recent League Lines item (August, 1974 QST) caused most nets to scrub "swap shop" sessions, but one subsequent interpretation seems to permit such activities providing they are scrupulously "non-commercial" and any given individual limits himself to "infrequent" participation (whatever that means)...

JA STATIONS TO GET NEW 80 METER "WINDOW" within the next month or so. Present JA allocations are 3500-3525 CW, 3525-3575 for phone; the new band is a 10-kHz slice from 3793-3803 kHz, will help JA-Europe and JA-U.S.A. QSOs.
Slow-Scan Enthusiasts can also look for a big increase in JA SSTV activity very shortly. Although over 100 stations have been licensed for slow-scan in Japan since April, 1973, reorganization of the Japanese Ministry of Post and Telecommunications has caused a big backlog to develop. Kinks are now being worked out, and a big influx of JA SSTVers is expected momentarily.



low-power single-band cw transceiver

Design and construction of a deluxe 1.5-watt QRP CW package for 20 meters

Wes Hayward, W7ZOI, 7700 SW Danielle Avenue, Beaverton, Oregon

There is little doubt that low-power (QRP) operation has become one of the more popular amateur radio activities in recent years. I got hooked on the QRP game a little over ten years ago when a need arose for light-weight portable gear for use on mountaineering trips. Although this goal was easily and quickly realized, my interest continues. Even today, after a few thousand QSOs with a power of only a few watts, I still get a thrill when another QRP contact is completed.

While some manufacturers are now seriously aiming products at the QRP market, the area is still ripe for the home experimenter. To many QRP enthusiasts, the only equipment which is considered for construction is that which is as simple as possible. While simplicity certainly has its merits, especially for portable operation in a severe environment,[1,2] home-station operation is greatly enhanced with equipment which is a bit more elaborate. This article describes a transceiver for 20-meter CW which is aimed at this improved performance.

One of the more significant deficiencies of many QRP stations is the lack of selectivity and dynamic range in the receiver. Although it is possible to obtain superb performance from a direct-conversion design, as demonstrated by the recent work of DeMaw,[3] there is still no substitute for a cleanly operating superhet. Hence, this approach was taken in this design.

Some of the simplicity of a direct-conversion design is retained by eliminating virtually all of the gain usually found at the intermediate frequency. The result is a receiver which is more than adequately sensitive and selective, but is still easy to duplicate. Provision is made for receiving both CW and ssb in the unit described, with sideband included mainly for use with vhf converters. Further simplification and reduced cost will result if one of the modes is deleted.

The transmitter portion of the package was designed with a number of objectives in mind. First, full transceive operation was desired. However, it was not acceptable to sacrifice the cleanliness of the system output. Experience with an earlier 40-meter transceiver demonstrated that this objective is not easily met with a casual design. The unit described in this article has been evaluated with lab quality test equipment, and all spurious outputs were more than 50-dB below the desired 1.5-watt output.

Additional design criterion for the transmitter included the desire for a built-in sidetone oscillator and a semi-break-in keying system. These features are not absolutely necessary, but are easily realized and add measureably to the operating enjoyment derived from the station.

Shown in **fig. 1** is a block diagram of the transceiver. The usual 9-MHz i-f and 5-MHz vfo scheme is used, providing good performance in a single-conversion system. A total of seven individual circuit boards are used, most of them based upon a simplified double-sided PC board technique outlined earlier. The use of double-sided board is highly recommended since it provides the low-impedance ground paths required for clean, spur-free performance.

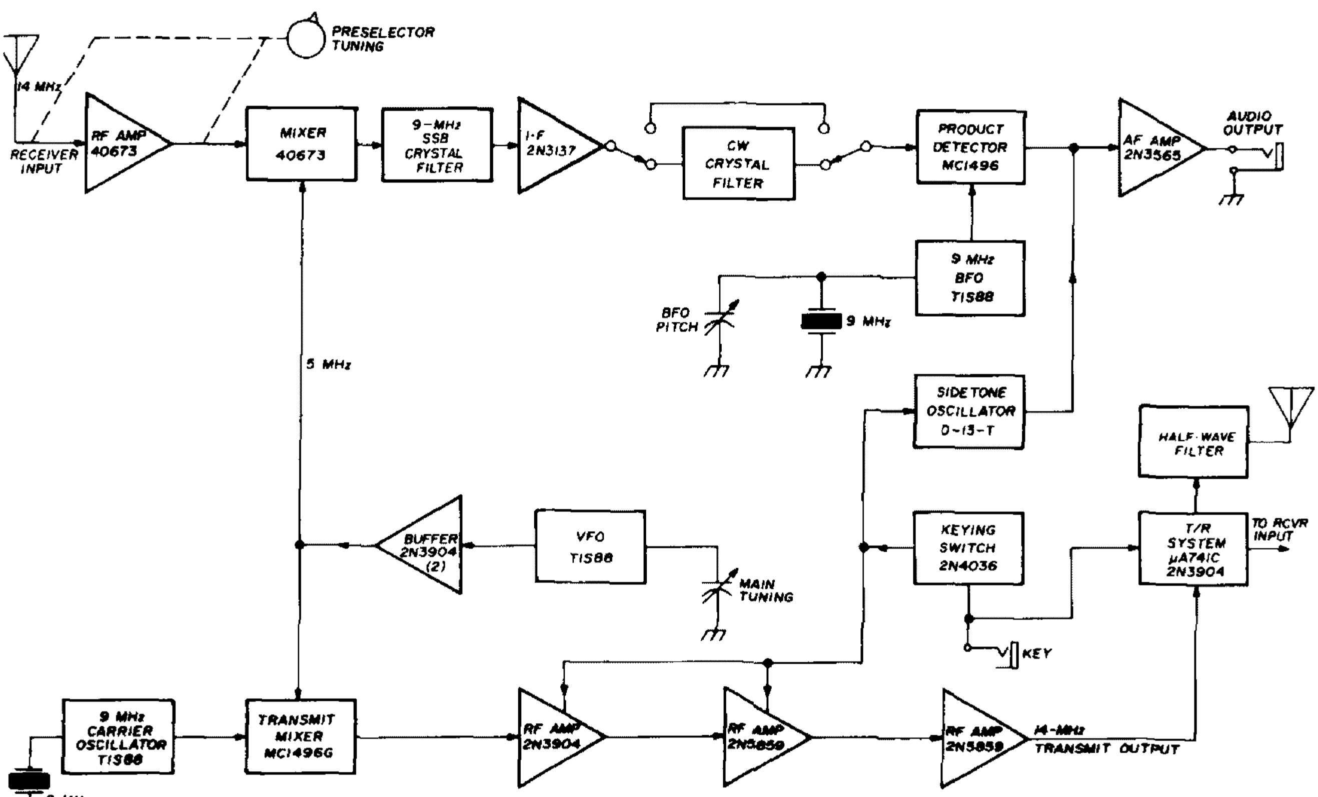


fig. 1. Block diagram of the deluxe QRP CW transceiver for 20 meters.

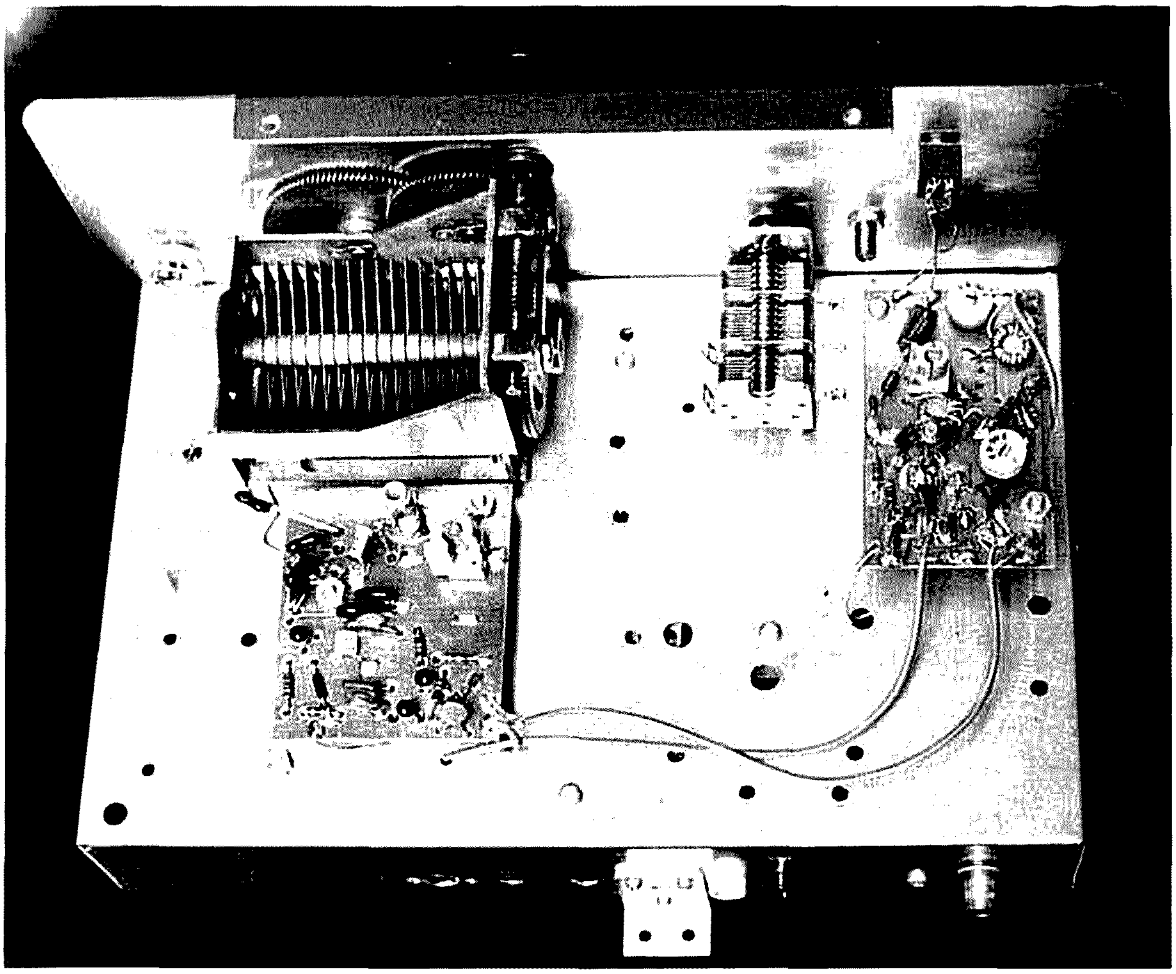
vfo design

Shown in **fig. 2** is the variable-frequency oscillator which controls both the receiver and the transmitter. The design offered several years ago by Hanchett[4] was used, although the original mosfet was replaced with a junction fet. I keep coming back to this design since it is

both stable and reliable. As with most oscillators, the selection of components is a large part of achieving stability. The inductor is wound on a ceramic form which was originally tuned with a powdered-iron slug. However, the slug was removed to enhance stability, and the number of turns on the coil was pruned to obtain the proper inductance. The vfo is tuned with the ever reliable and easy-to-use capacitor from a surplus ARC-5 Command transmitter. Although becoming scarce, these capacitors can still be found in junk boxes and at hamfest flea markets. With the components shown, the oscillator tunes from 5.0 to 5.55 MHz.

A number of commonly available field-effect transistors can be used in this oscillator. I used a TIS88 which is a plastic device very similar to the popular 2N4416. The Motorola MPF102 also performs well in this circuit. The oscillator supply voltage is stabilized with a 6.8-volt zener diode. The voltage rating here is not critical.

The vfo is buffered with a single-stage feedback amplifier using a pair of 2N3904 transistors. Again, transistor type is not critical in this application, although devices with an f_T of at least 250 MHz should be used. The output of the buffer is 3 volts, peak-to-peak and sinusoidal. If it is suspected that significant harmonic energy might be present in the oscillator output, a lowpass filter could be included.[5]



Top view of the 20-meter CW transceiver, showing the vfo (left) and receiver rf amplifier (right).

receiver front-end and i-f filtering

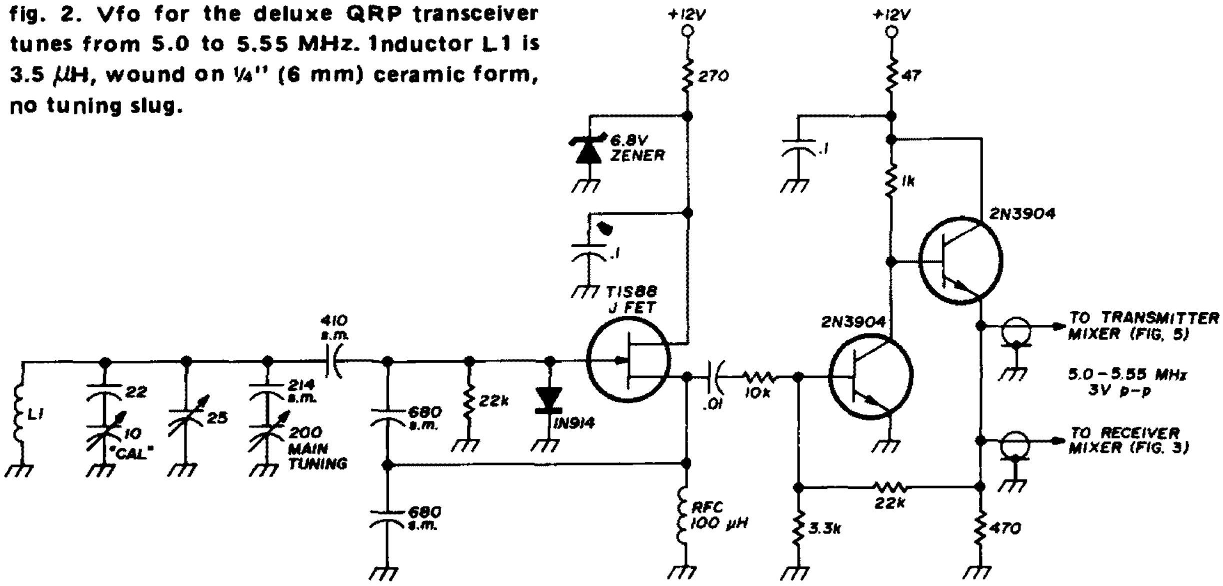
Presented in fig. 3 is the front-end and filter section of the receiver. For the most part, the design is quite standard and is easily duplicated. Both the rf amplifier

and the mixer circuits use RCA 40673 dual-gate mosfets. Since my transceiver tunes a 500-kHz range, a dual-section variable capacitor was included for front-panel preselector tuning. An earlier version of this transceiver tuned only the CW portion of the band, and front-panel tuning was not needed. The toroid cores used in the receiver and in the transmitter section described later are very similar in

on the respective coils. The minimal degradation in gain and noise figure should present no problem in typical applications. The dynamic range of this receiver has not yet been measured.

A coarse rf gain control is provided by a front-panel switch which decreases the gate-2 bias from the nominal 4-volt level to ground. This yields a gain reduction of about 15 dB. Since the application of

fig. 2. Vfo for the deluxe QRP transceiver tunes from 5.0 to 5.55 MHz. Inductor L1 is 3.5 μH, wound on ¼" (6 mm) ceramic form, no tuning slug.



inductance and Q to the Amidon T-50-6.* Substitution of Amidon cores should be possible using the same number of turns shown.

Sensitivity measurements revealed a system noise figure of around 7 dB. Hence, the receiver is probably quite a bit hotter than can be used in normal locations. Much of the dynamic range possible with mosfets was retained by terminating the mixer output with a 330-ohm resistor. Since drain nonlinearity is the usual mechanism for blocking in a fet mixer, a decrease in output load impedance results in relative freedom from this problem. Similarly, the load resistance presented to the drain of the rf amplifier is about 1000 ohms. If dynamic range problems are encountered, further improvement would result if the gate of each stage were driven from a tap

gain reduction can often decrease the immunity of an fet rf amplifier to cross-modulation and IMD, passive front-end signal attenuation would be a better means of gain control.

The output of the mixer is routed through a coax cable to the first crystal filter. This filter is always in the signal path and is used for ssb reception. The filter I chose was the WF-8 model manufactured by Wheatlands Electronics.† This eight-pole unit performs well in this application, and probably represents one of the better component buys around.

The output of the first filter is applied to a low-gain amplifier. A junk box 2N3137 was used, although this stage is not critical, and could probably be replaced with a 2N3904 or a 2N4124. Alternately, if some impedance matching is done at the input, a dual-gate mosfet

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

†Wheatlands Electronics, P.O. Box 343, Arkansas City, Kansas 67005.

with a 510-ohm drain load resistor should perform well in this slot. The main function of this stage is to provide a proper driving impedance for the second filter with a minimum gain.

During CW operation, a KVG type XF-9M four-pole crystal filter is switched into the system.* The switching is done with a pair of inexpensive slide switches which are mounted on the circuit board.

input resistors have been chosen to properly terminate the KVG crystal filter. The bfo is a simple fet crystal oscillator which is trimmed from the front panel with an 80-pF variable capacitor.

The audio section consists of a pair of 2N3565 transistors and provides around 70-dB of gain to drive high-impedance headphones. This amplifier is built on the same board that contains the sidetone

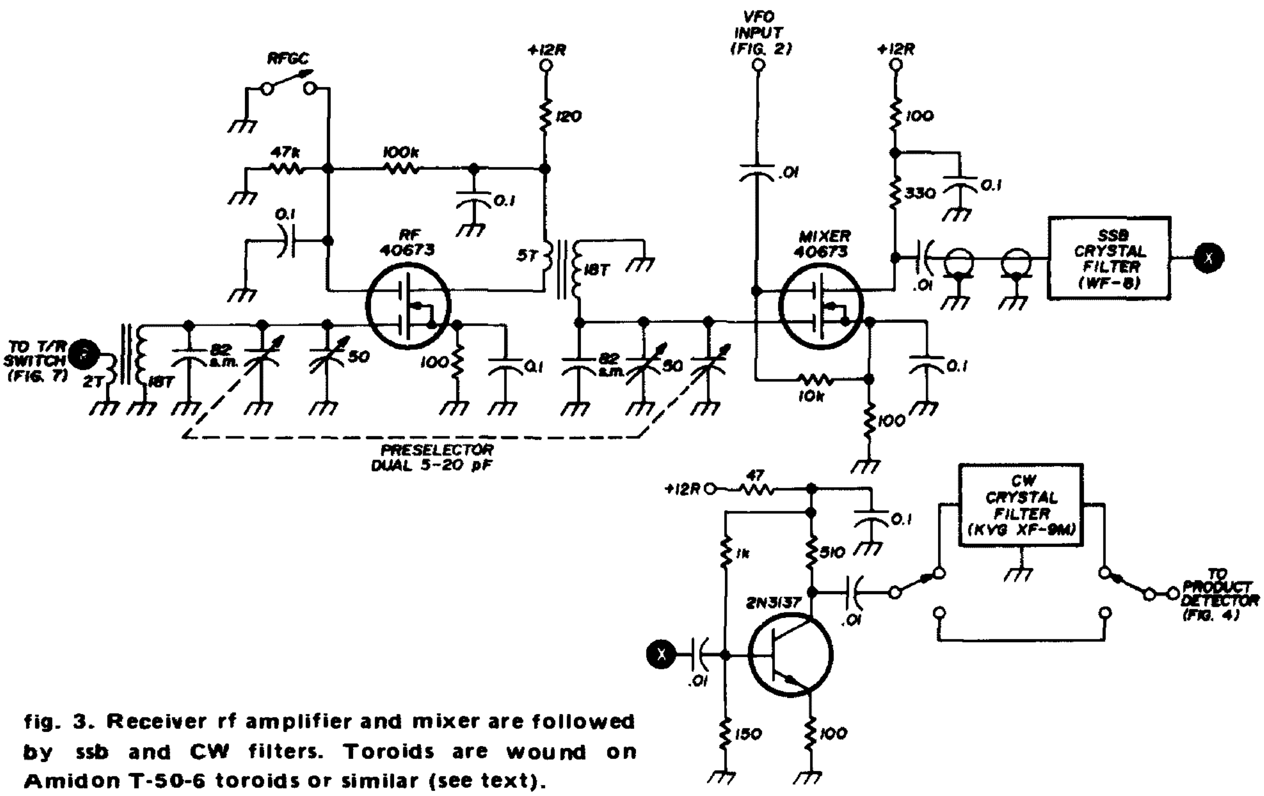


fig. 3. Receiver rf amplifier and mixer are followed by ssb and CW filters. Toroids are wound on Amidon T-50-6 toroids or similar (see text).

The switches are ganged together by drilling small holes in the plastic handles and attaching a strip of scrap PC board. A spade lug is also bolted to the connecting strap. A tapped, ¼-inch (6-mm) spacer is then screwed onto the spade lug. This spacer extends through the front panel where a knob is mounted.

product detector and audio

Shown in **fig. 4** is the product detector and bfo used in the receiver. The detector uses a Motorola MC1496G integrated circuit. The configuration is generally the same as that shown in the Motorola applications literature, except that the

*KVG crystal filters are available from Spectrum International, Box 1084, Concord, Massachusetts 01742.

oscillator and control circuitry and is shown in **fig. 7**.

Although the lack of gain at the intermediate frequency makes the possibility of adding agc a bit difficult, this simplified distribution does have its virtues. The main advantage is that product detection occurs at fairly low signal levels. This minimizes the noise modulation effects which often occur in product detectors. Another problem which is avoided is the effect of bfo leakage. Often, stray bfo energy finds its way into a high gain i-f amplifier, causing both intermodulation and noise modulation to occur. This is avoided in a system of this kind. The overall result is a receiver which sounds exceptionally crisp and clean, a virtue usually limited to direct-conversion

receivers and some well designed superhets.

transmitter mixer

Shown in fig. 5 is the transmitting mixer and carrier oscillator for the transceiver. Although it would be possible to replace the carrier oscillator with some energy from the bfo, this would necessitate the introduction of some offset of

Motorola chip is by far the more satisfactory and is easily applied.

Balance is maintained in the mixer by using a center-tapped tuned circuit in the output. This is realized by putting a bifilar winding of 15 turns on a toroid core and tuning the series combination. The required power-supply voltage is injected on the center tap and output is extracted from the tuned circuit with a

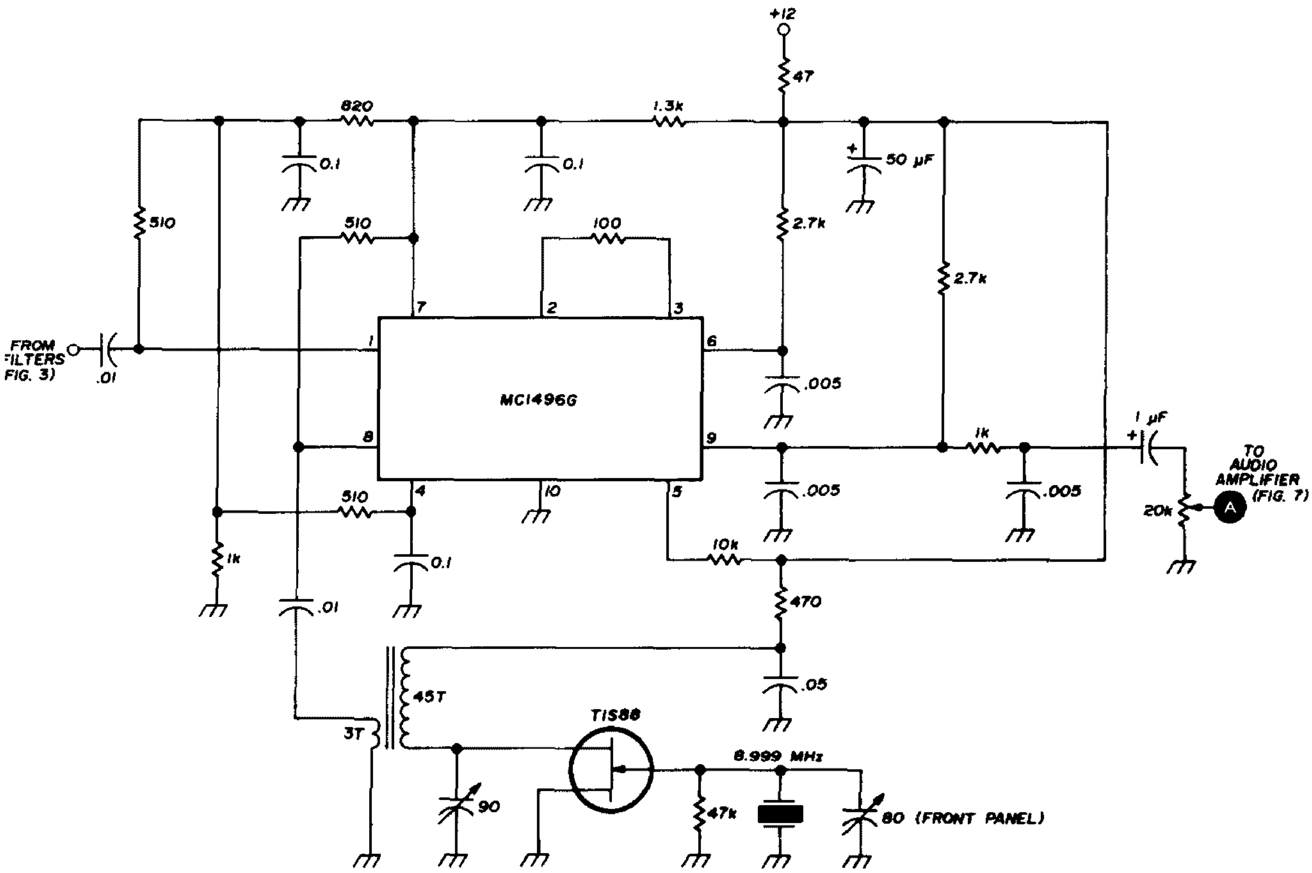


fig. 4. Product detector and bfo for the receiver used in the QRP transceiver. Transformer is wound on Amidon T-50-6 toroid or similar.

the main system vfo during transmit. The use of a separate crystal oscillator was considered to be the simpler solution.

The transmitting mixer itself uses another Motorola MC1496G doubly-balanced modulator IC. This chip is ideally suited for this application due to the excellent balance available. This significantly reduces the amplitudes of many of the spurious products in the output from those which would appear in the output of a single-ended mixer. I have also used the RCA CA3028A and diode rings in this application, although the

3-turn link. An alternate output network would be realized by shunt-feeding the MC1496G output collectors with rf chokes. Then, the high-impedance end of a tuned circuit could be lightly coupled to one of the output terminals. This method would have the advantage of being more easily bandswitched.

The input levels feeding the mixer IC are not extremely critical, although severe overdriving should be avoided since this will cause some deterioration in the rejection of spurious output products. In the mixer shown, the carrier port is driven

with a little over 0.5 volt, rms, and the signal port has about 200 millivolts of drive. During receive periods, the supply voltage is removed from the mixer/carrier-oscillator board.

transmitter power chain

The three-stage power amplifier which completes the transmitter is shown in fig. 7. The 2N3904 pre-driver and the gain might be provided by a single stage. First, I have found that it is usually worthwhile to add an additional stage with a decreased gain-per-stage to insure stability. The cost increase is minimal, but the unconditional stability is quite assuring.

Class-A operation has the advantage of preserving linearity. This is somewhat important, even in a CW rig. In previous

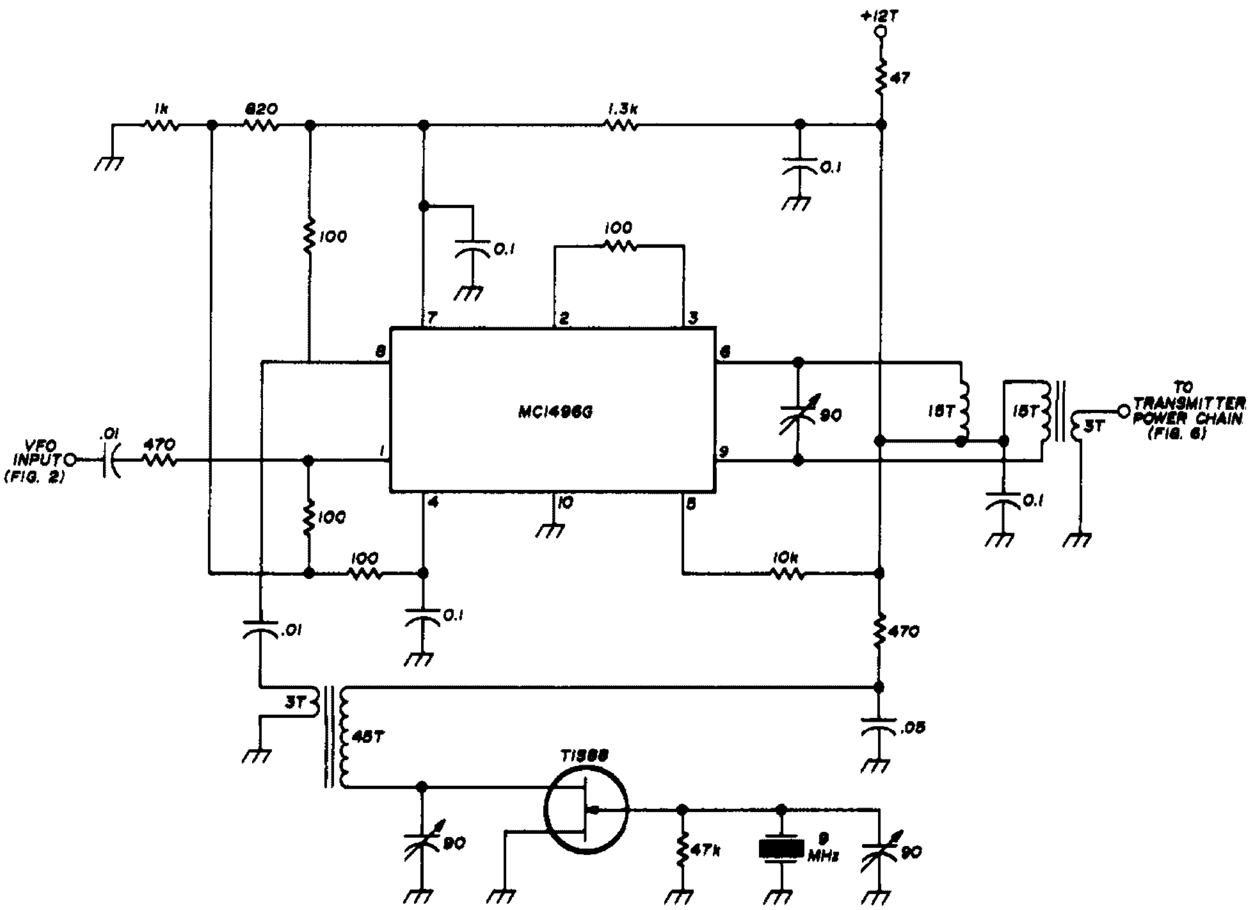


fig. 5. Transmit mixer and carrier-oscillator circuits. The frequency of the 9-MHz crystal oscillator is adjusted to the center of the 9-MHz i-f with the crystal filter (fig. 3) switched in. Transformers are wound on Amidon T-50-6 toroids or similar.

2N5859 driver are keyed. The final amplifier also uses a 2N5859. This Motorola TO-5 device is an excellent, general-purpose QRP device with a price tag of under one dollar. Since it has an f_T of around 250 MHz, it should be usable up as high as the six-meter band, with power output up to about two watts.

Both the pre-driver and the driver stages are operated as class-A amplifiers with emitter degeneration. There are a couple of reasons for this, even though sufficient transmitters it was noted that the spurious components in the output would be increased if an amplifier was allowed to saturate. The reason is that saturation would occur for the primary driving frequency, but the amplifier would still behave in a fairly linear fashion for spurious products. The net effect was that much of the filtering prior to the amplifier was negated.

The bandwidth of the amplifier strip of fig. 6 is rather restricted. With the

system aligned at 14.065 MHz, the commonly used QRP frequency on 20 meters, the output was down by about 30 dB when the vfo was tuned to the middle of the phone band. Most of the output power could be obtained, however, by retuning the output of the pre-driver. If operation of the transmitter over the total 20-meter band is required, the builder should provide for front-panel adjustment of this tuned circuit. As shown, the system is flat over the CW portion of the band.

The output power of the transmitter was measured at 1.5 watt into a 50-ohm load. When the output of the transmitter board was investigated with a lab-quality spectrum analyzer, it was found that the second harmonic was down by only 27 dB. However, all other spurious outputs were over 60 dB down. Additional harmonic rejection is easily obtained with the half-wave filter at the transceiver antenna terminal. With two class-A stages being keyed, the backwave was more than 70 dB below the usual output.

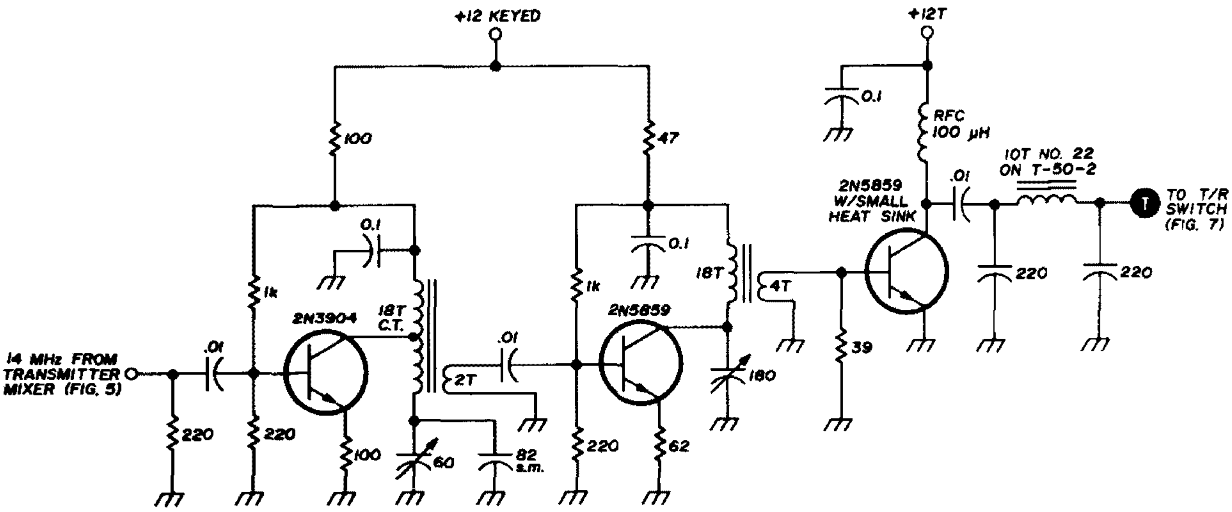


fig. 6. Transmitter power chain provides 1.5-watt output into a 50-ohm load.

control circuits

Shown in **fig.** 7 is the final board which completes the transceiver. This board contains not only the audio system mentioned earlier but the sidetone oscillator, a keying transistor and the T/R circuitry. The sidetone is obtained from a simple relaxation oscillator using a programmable unijunction transistor (PUT). The sawtooth output is attenuated and applied to the input of the audio amplifier. Transmitter keying is accomplished with a series switch using a 2N4036 silicon TO-5 transistor. Almost any silicon pnp device can be used in this slot.

In the circuits discussed above, the power supplies on the schematics have been labeled as +12V, +12R or +12T. All circuits labeled with +12V have voltage applied at all times. However, those with the R or T suffix have power applied only during receive or transmit intervals, respectively. The two voltages are derived from the antenna relay, K1. I used a dpdt relay with an 800-ohm, 12-volt coil, with the second set of contacts switching the antenna.

You may have noted that the product detector has power applied at all times, even during transmit intervals. Initially, the detector was muted during transmit. However, an objectionable click appeared when the receiver came on again. This was the result of the large current surge in the 50-μF decoupling capacitor used on the detector board.

The transmit-receiver logic is based upon a μA741C operational amplifier IC used as a differential comparitor. Under key-up conditions, the 2.2-μF capacitor is fully charged. However, when the key is

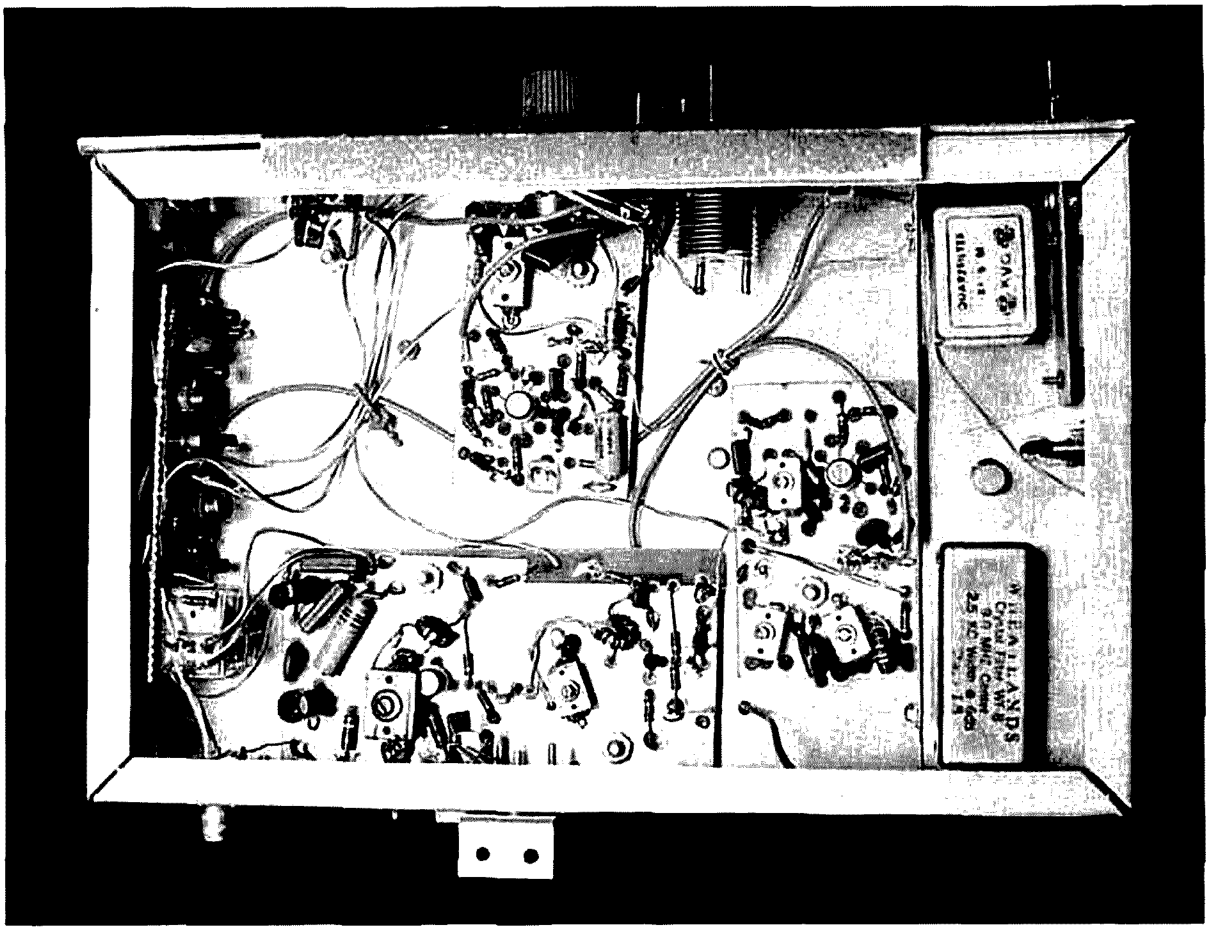
closed, this capacitor is discharged. When this happens, the voltage on the inverting input of the op amp (pin 2) drops below half of the power supply potential. This causes the op amp output to switch to a high state, near the 12-volt supply, and saturates the 2N3904 relay driver. When the key is released again, the timing capacitor begins to charge toward the 12-volt supply through the 220-kilohm resistor. When the capacitor passes the 6-volt point, the op amp changes state again and the relay opens. The hold-in time can be varied from the 0.5 second I used by changing the 220-kilohm timing resistor.

There are two keying inputs shown. One is for the usual hand key. The other uses a *two-circuit* phone jack, with +12 volts available on the second pin to supply power to an external electronic keyer. This keyer[6] is based upon differential comparitor logic similar to that used for the T/R control.

additional thoughts

The transceiver described and shown in the photographs was packaged in an LMB type CO-2 cabinet. The extra holes shown in the chassis are a reminder of an earlier receiver which resided in the same enclosure. Although there is an abundance of extra room left, this will eventually be used for a variety of accessories, including equipment for at least one vhf band.

An obvious extension of this design would be the addition of other bands. Of the possible combinations, the easiest would be a 20- and 80-meter CW transceiver. The receiver can be made operative on 80 meters merely by switching the tuned circuits in the front end.



There's plenty of room under the chassis. Crystal filters are at right, next to transmit mixer and carrier oscillator. Transmit power chain is at bottom with control, keying and audio circuits to the left.

The transmitter mixer is moved to 80 meters by changing the output network in the transmitting mixer as described earlier. The output of the mixer should then be well filtered for the band in use. In this case, a lowpass filter would be suitable for 3.5 MHz with a bandpass filter being switched in for 14-MHz operation.

The transmitter power chain could be replaced with a broadband design with appropriate lowpass filters switched into beyond the relatively simple systems considered in this article.

The performance of this transceiver has been more than satisfactory. Using only a ground-plane antenna, contacts have been made all over the United States and Canada as well as with VK, JA, UAØ and DM. While the unit does not represent the absolute ultimate in simplicity, the superior performance is well worth the minimal extra effort and expense.

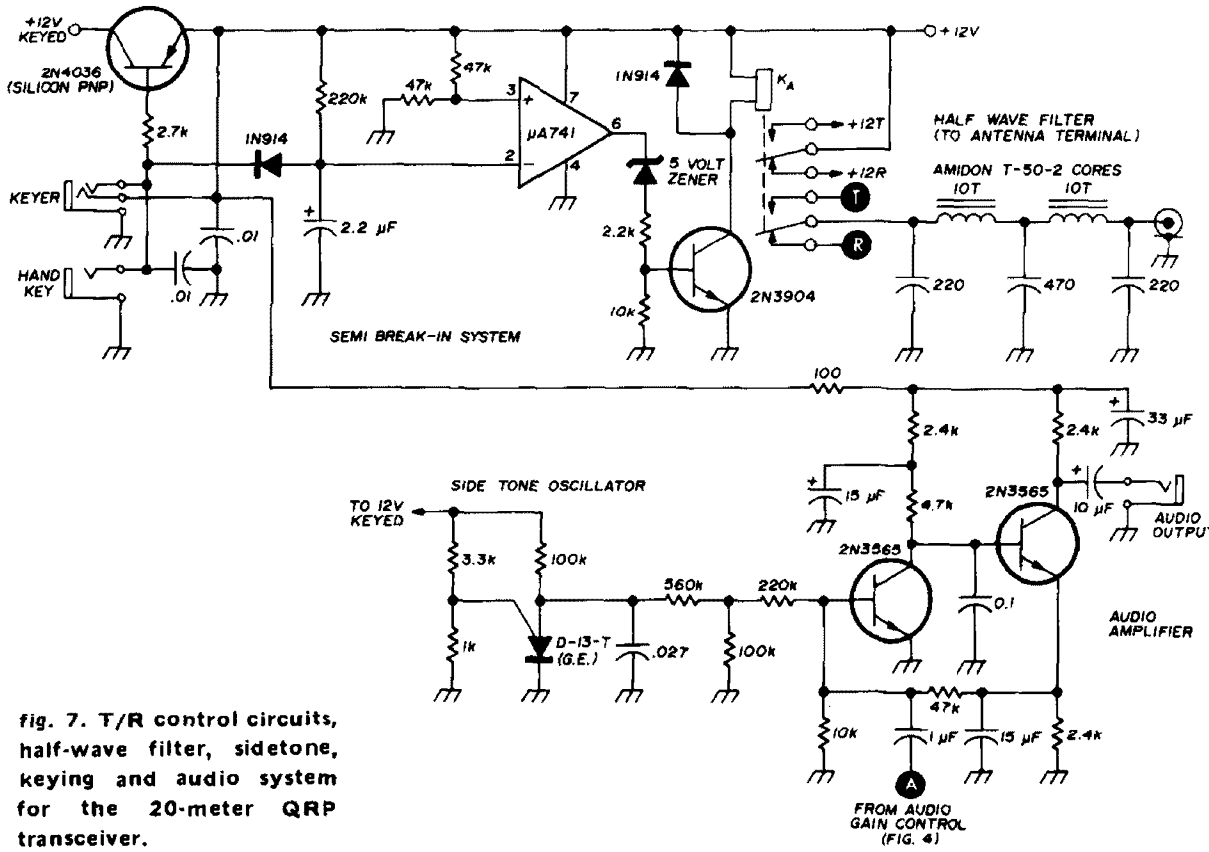


fig. 7. T/R control circuits, half-wave filter, sidetone, keying and audio system for the 20-meter QRP transceiver.

the output. If such an approach is taken, it is important that the filtering between the mixer and power amplifier be sufficient to provide an output free of spurious products. In most cases, a two- or three-section bandpass filter will do the job.

If bands other than 80 and 20 meters are considered, it will be necessary to derive other local-oscillator signals. This could be achieved either by bandswitching the vfo or by premixing the existing 5-MHz signal. Other, more elaborate synthesis techniques would yield superb performance, but they go

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ham radio

scattering characteristics of artificial radio aurora

Artificial aurora, created at will by high-power, high-frequency transmitters, may prove useful for long-distance vhf communications

Victor R. Frank, WB6KAP, Stanford Research Institute, Menlo Park, California 94025

The U.S. Department of Defense has recently disclosed that it is now possible for man to create his own artificial radio aurora (ARA) capable of scattering radio waves of frequencies up to 450 MHz. The ARA can be produced in either the ionospheric E- or F-regions over a source of very powerful high-frequency radio waves directed upwards. Don't, however, expect to see this "aurora." The optical effects of ionospheric modification are very weak, corresponding to less than a 30% change in the level of some characteristic lines in the night airglow spectrum.[1]

Research has taken place at three sites. One,[2] near Platteville, Colorado, operated by the Institute for Telecommunications Sciences (ITS), uses a transmitting system with an effective radiated power of 40 megawatts. Two transmitting arrays are used; one, for the frequency range 2.5 to 5 MHz is composed of five crossed dipoles, and the other, usable between 5 and 10 MHz, uses ten crossed dipoles. The crossed-dipole elements are fed in phase quadrature to produce circular polarization. It has been found that the ordinary magnetoionic component[3] produced by right-hand circular polarization (in the northern hemisphere) produces the strongest ARA. The second ionospheric heating facility at Arecibo, Puerto Rico, operated by Cornell University, uses a transmitter with one tenth the power output and an antenna with 10 dB more gain (the 1000-foot dish). A third facility, located at Gorki, 400 km east of Moscow, USSR, uses a transmitter with 60-kW average power and 22-dB antenna gain.[4]

In the ionosphere, just below the reflection height, the high-frequency wave is retarded and energy is imparted to the electron gas, raising its temperature. It is forced to expand along the magnetic field lines, creating field-aligned ionization irregularities. Although the fractional change in ionization density through these irregularities may be only a percent or so, each irregularity scatters coherently. Many of these long, thin irregularities are capable of scattering high-frequency and vhf signals just as the natural aurora does. The primary differences as far as the radio amateur is concerned are that this is a very localized aurora, it may be turned on and off within seconds, and its height is determined by the frequency of the heating transmitter.

Since the scattering region is localized

and the Doppler spreading is not great, voice signals reflected by ARA are still intelligible although there is rapid and deep flutter fading.

A companion article appearing in *QST*[5] this month describes ARA sounding and communications experiments and the directional properties of the scatterers. Since I will go into further detail here, it may be to the reader's advantage to read the *QST* article first.

F-region scatter observed

Soon after the announcement of the triggering of spread F by the Platteville facility,[6] experimenters noted echoes on high-frequency sounding paths that were determined to be due to reflection from field-aligned irregularities from the F-region over Platteville. Fig. 1 shows a chart recording made in March, 1971, at an SRI field site in central California of experimental transmissions from a station in Arkansas. The frequency was chosen on the basis of oblique soundings to be a few MHz above the direct path MOF. A clear cause-and-effect relationship is shown, with the 20-MHz signal fading into the noise some 20 seconds after turnoff of the ionospheric modification transmitter. A-m voice modulation on this signal was quite readable.

The fading rate and the lengths of time that the scattered signal remained after turnoff of the heating transmitter varied with the frequency and with the scattering angle. The lower frequencies and the forward scatter paths had lower fading rates and were more persistent than higher frequencies and backscatter paths.

High-frequency soundings had indicated that even higher frequencies might be usable, so equipment for receiving TV video carriers was set up at another field site near Bakersfield, California. The receiving system used an array of eight commercial LPA antennas (fig. 2), crystal-controlled converters and narrowband receivers with recording on chart and tape. Fig. 3 shows chart recordings of some of the signals observed while the Platteville transmitter was being operated on a one-minute on, one-minute off cycle.

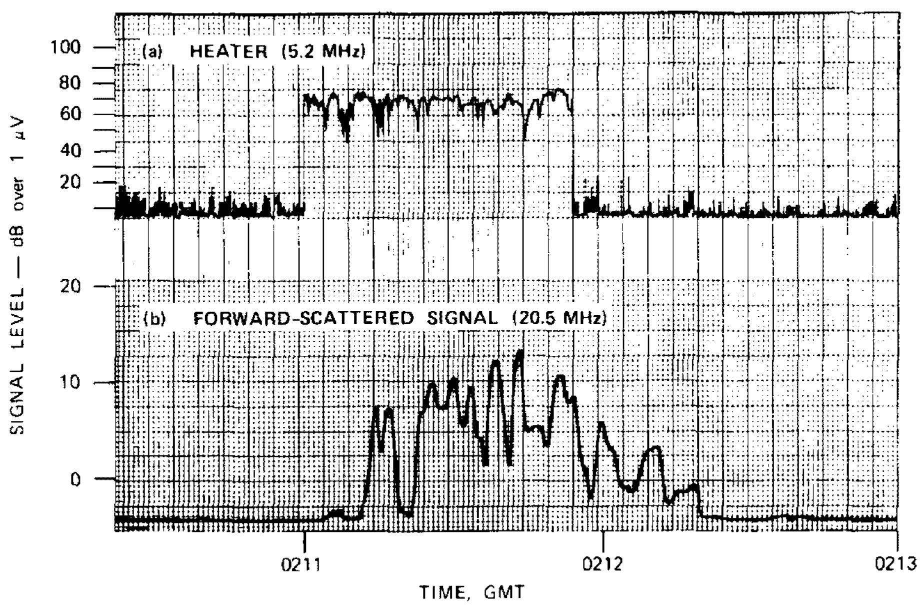

fig. 1. Field-strength recordings made in central California of the ITS ionospheric modification transmissions (A) and a transmission from Arkansas at a frequency above the direct path MOF (B).

It is difficult to identify a television station solely on the basis of its video carrier frequency.* Signals were never strong enough to demodulate and, in addition, too much co-channel interference existed from stations in northern and southern California. Based on ray-

*In the United States, Canada and Mexico, vhf TV stations are assigned video carrier frequencies of either 1.24, 1.25 or 1.26 MHz above the lower edge of the channel. These are referred to as minus-, zero- or plus-offset, respectively.

tracing calculations, these TV signals are believed to have originated from stations in northern Mexico and southwest Texas.

E-region scatter

One of the most recent developments in ionospheric modification has been the observation of artificial radio aurora in the E-region over Platteville, Colorado. This development was made possible by the addition of five low-frequency crossed dipoles in an array at Platteville, which allowed operation at frequencies as low as 2.8 MHz. These frequencies are reflected from the E-layer during the daytime.

Actually, most amateur experience with natural field-aligned scattering is that from the E-region, since the aurora rarely penetrates far enough south for echoes to be obtained from the F-region. Fig. 4 shows the intersection with the earth of scattering cones of various angles to the magnetic field at a height of 110 km over Platteville, Colorado. (A similar picture for F-region reflectors is shown in the companion *QST* article.) Specular scattering may take place between stations located on intersections of supplementary cone angles (their sum equals 180°). These curves were derived for a single field-aligned scatterer. Actually, such scatterers are present in a scattering region of 100-km diameter and 10-km thickness. This distribution will spread the coverage by a similar distance. Soundings were made over a path from a transmitter site south of Albuquerque, New Mexico, to an ITS receiving site near Haswell, Colorado, where there is a 60-foot parabolic dish antenna (fig. 5). One of the sweep frequency soundings taken over this path, fig. 6, shows time delay, frequency coverage and radar cross section versus frequency. The sounding transmitter was turned off during those portions of the frequency sweep that fell within TV channels 4, 5 and 7.

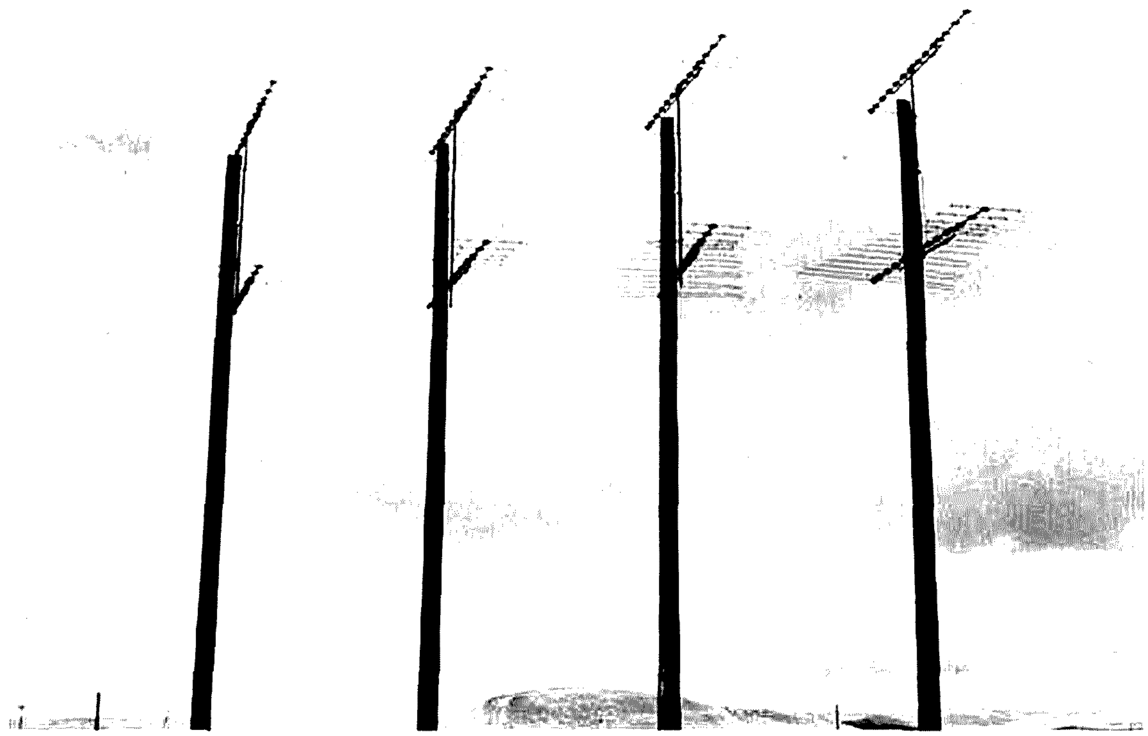

fig. 2. Array of log-periodic TV antennas used at site near Bakersfield, California, to monitor TV video carrier signals scattered by ARA.

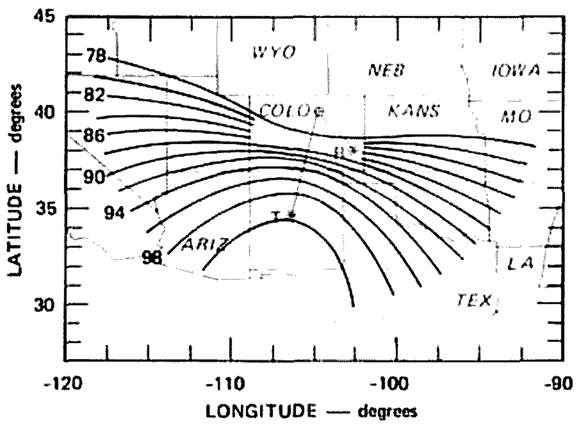

fig. 4. Contours of intersections with the earth of scattering cones from field-aligned scatterer at 110 km over Platteville, Colorado. Communication can be established between stations located on contours of supplementary aspect angles.

Radar cross section may be a new concept to some of you. It is here defined by the radar formula

$$\sigma = \frac{P_r\,(4\pi)^3\,r_1{}^2 r_2{}^2}{P_t G_t G_r \lambda^2}$$

where P_t is the transmitted power output, P_r is the received power input, λ is the wavelength and r_1 and r_2 are the distances from transmitter and receiver to target, respectively, in meters, and G_t and G_r are the transmitting and receiving antenna power gains over an isotropic radiator. This value of received power, P_r, must compete with the noise power, P_n which is

$$P_n = FkTB$$

where F is the ratio of cosmic noise power to that produced by a termination (Johnson noise), k is Boltzmann's constant (1.38×10^{-23}), T is the absolute temperature in degrees Kelvin, and B is the receiver bandwidth in Hz. To spare the reader the exercise of going through the numbers, I have calculated the minimum detectable cross section for an amateur CW station in various amateur bands having the following characteristics:

P_t = 500 watts
r_1 = r_2 = 10^6 meters
B = 100 Hz
P_r/P_n = 1
G_t = G_r = 10 (10 dB) 28 MHz
= 20 (13 dB) 50 MHz
= 40 (16 dB) 144, 220 MHz
= 60 (18 dB) 432 MHz

The minimum observable cross sections are about 10^3 to 10^4 square meters (30 dBsm to 40 dBsm). These are shown in fig. 7 along with the range of cross sections observed from the Platteville heated region under ideal conditions for both E- and F-region heating. You can see

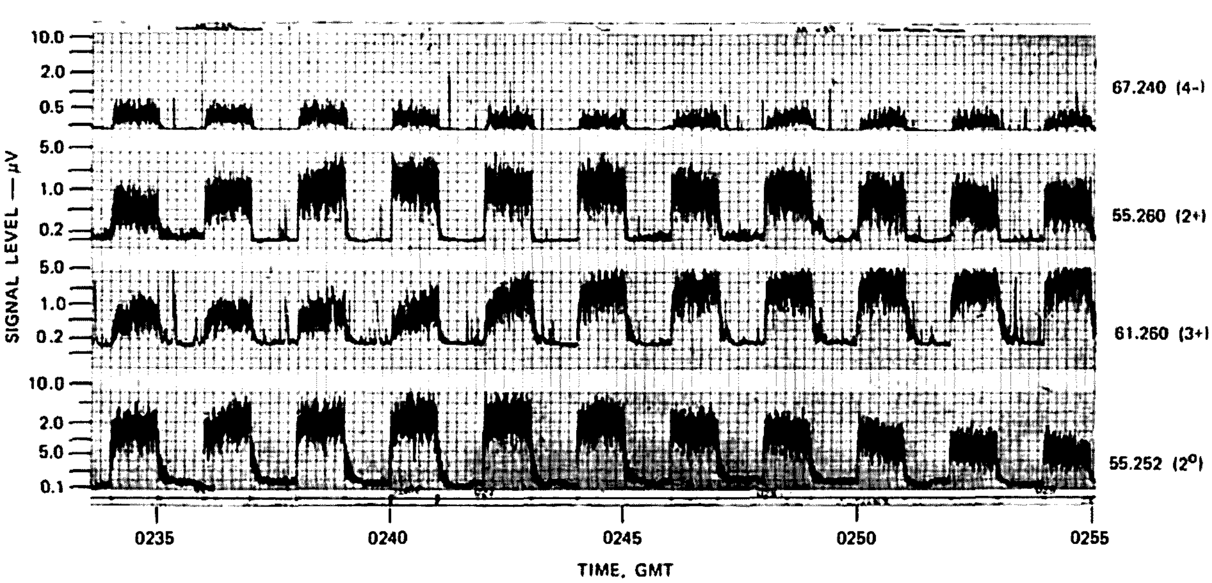

fig. 3. Field-strength recordings of TV signals received during 1-minute on, 1-minute off cycling of ionospheric modification transmitter.

that detection at 50 and 144 MHz is well within the capabilities of the advanced amateur station.

It is quite possible that some amateurs may have already communicated two-way via ARA without realizing they were doing so. During the last test series at Platteville, about 264 hours of heating experiments were conducted (between September 10 and November 2, 1973). Of these, 47 were in the prime evening time interval, 0000-0600 UT. It is hoped that radio amateurs will have a chance to participate in future tests presently scheduled for Platteville, Colorado, some time during the first half of 1975 and for Arecibo, Puerto Rico, during the first weeks of April, 1975.

ionospheric modification at arecibo

Since the Arecibo heating antenna beam is narrower than that at Platteville, the expected cross sections are some 10 dB lower. Also, since Arecibo is not licensed at present for frequencies below 5 MHz, E-region heating is not possible and F-region heating is possible only during daylight and early evening hours. Fig. 8 shows a map of intersections with the earth of scattering cones from a field-aligned scatterer at 300-km height over Arecibo, Puerto Rico.

Most likely to benefit from Arecibo ARA are the Caribbean, Venezuela, Columbia and possibly Central America—areas with very little amateur vhf activity.

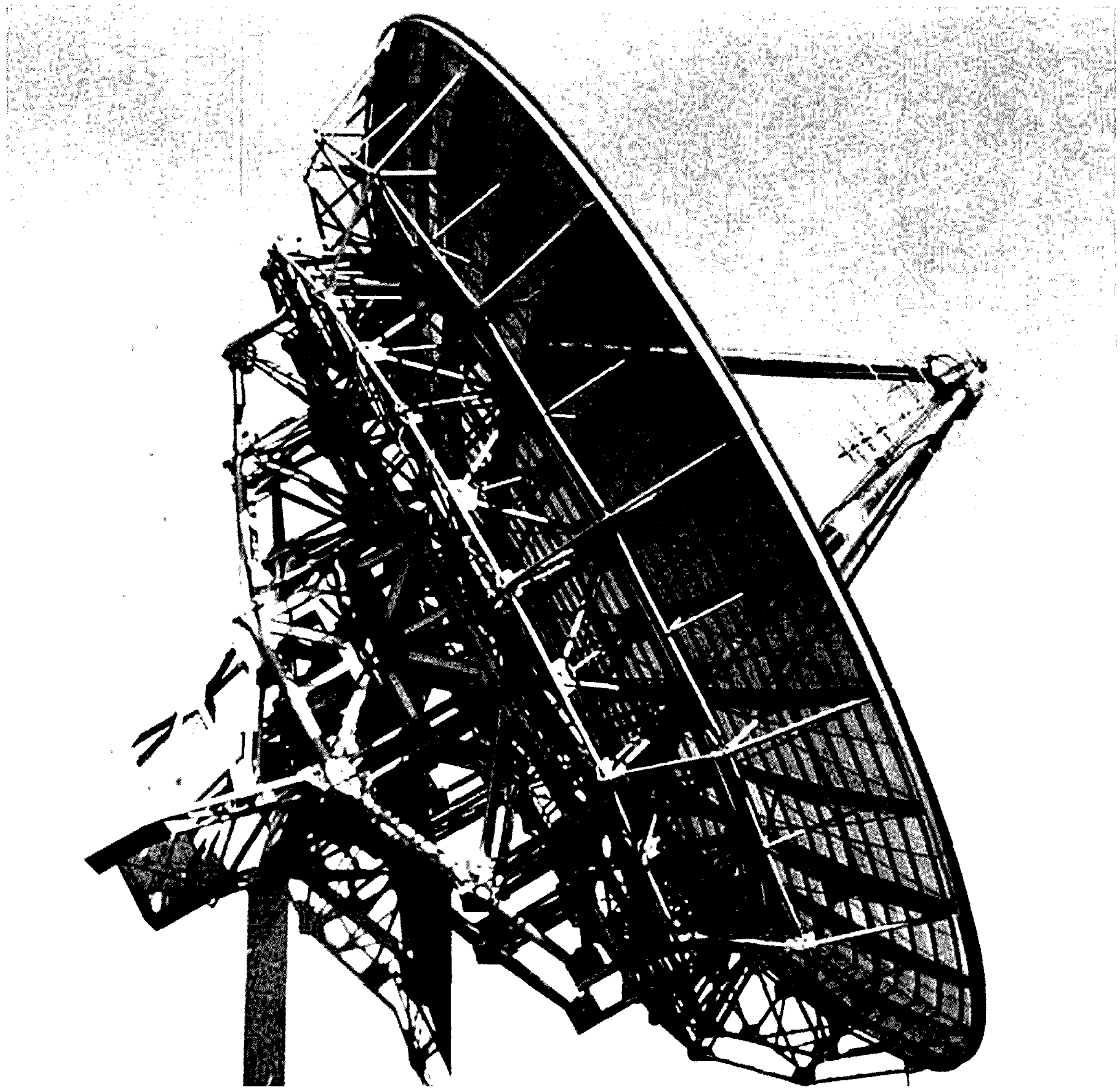

fig. 5. Sixty-foot dish and LPA feed at ITS field site near Haswell, Colorado.

Perhaps some vhf DXpeditions would be in order. These are good areas, and April is a good time of year for transequatorial scatter.

other possibilities

If you are not fortunate enough to be in the ARA coverage areas of Platteville or Arecibo, do not despair. Much remains to be learned about what is really happening during ionospheric modification. There is another form of scattering caused by plasma and ion-acoustic waves[7,8] which, though very weak, may be usable for stations within line-of-site of the ARA. Amateurs, being spread all over the country, are in a unique position to investigate paths not previously available to researchers.

In addition, much of what we have learned during these experiments will be of use to the radio amateur interested in scatter and auroral communication. Field-aligned ionization in nature? Yes, even at mid-latitudes, if you know how to look for it.[9] Further information on coverage areas for scattering by the natural aurora may be found in an article by R.L. Leadabrand.[10] I am convinced that the so-called *X-mode* (50-MHz signals backscattered from sporadic E-patches) is due to field-aligned ionization in those patches.[11] And transequatorial propaga-

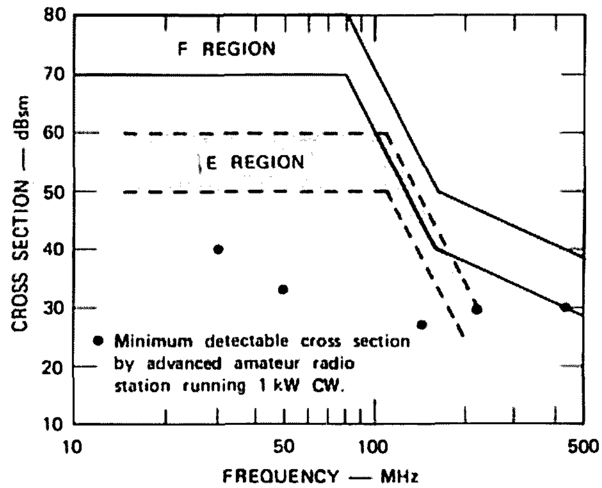

fig. 7. Radar cross-section of the Platteville ARA versus frequency for E- and F-layer echoes.

tion is also probably due to forward scattering from field-aligned irregularities in the equatorial F-region.[12] In addition, experimenters have found that meteor-trail debris becomes field-aligned.[13,14]

Ranging capability, something lacking in most radio amateur stations, would prove invaluable in identifying these propagation modes. It is possible some may have been used for decades without proper identification. Ranging capability implies pulse or linear sweep-frequency CW operation, neither of which is in amateur use in the vhf bands. But here, perhaps, is an opportunity for the radio amateur to use some of those radar

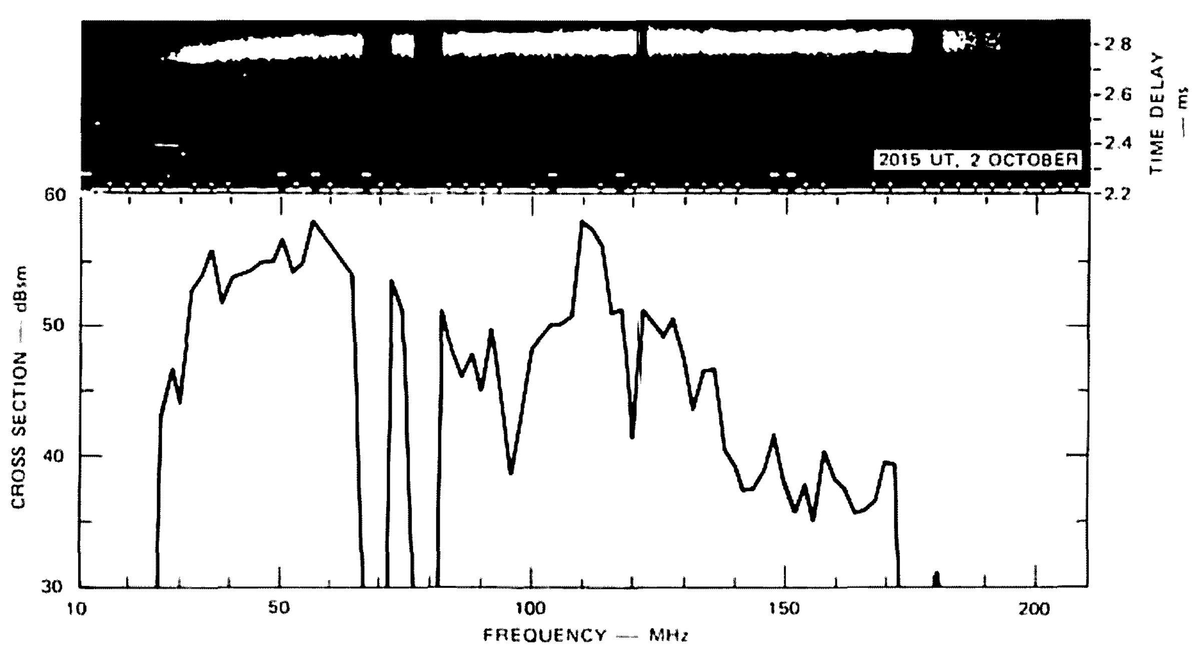

fig. 6. Sweep-frequency sounding received at Haswell, Colorado, showing echo from E-region ARA.

signals in the 220- and 420-MHz bands for scientific investigation.

acknowledgements

I wish to thank W.F. Utlaut and his associates at the Institute of Telecommunications Sciences, Boulder, Colorado, for their cooperation in the conduct of these experiments and for graciously allowing us to use their Haswell facility. Acknowledgment should also be made to the other organizations that were part of the team that explored the scattering properties of ARA. They are Raytheon Corporation, Sudbury, Massachusetts; Riverside Research Institute, New York City, New York; Barry Research, Palo Alto, California; and the Aeronomy Corporation, Champaign, Illinois. This research was sponsored by the Defense Advanced Research Projects Agency through the Office of Naval Research.

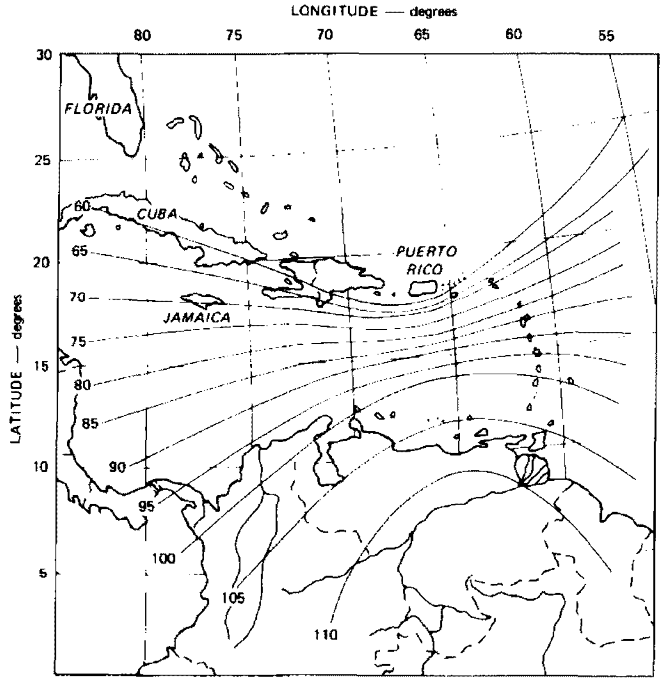

fig. 8. Contours of intersections with the earth of scattering cones from field-aligned scatterers at 300 km over Arecibo, Puerto Rico. The contour values are the angles between the cones and the geomagnetic field.

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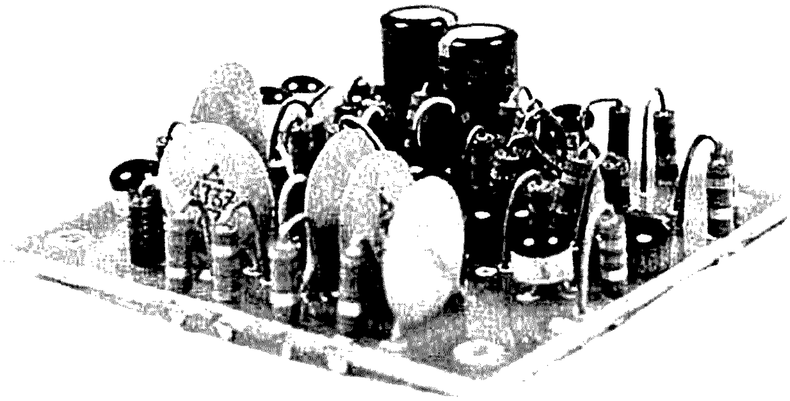


improved channel scanner for vhf fm

A complete four-channel vhf fm scanner that can be built for ten dollars

Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14612

If someone were to tell you that an integrated-circuit receiver scanner could be improved by redesigning it with discrete components, you'd say they were crazy! But, forced with difficulties in obtaining TTL logic elements for the K2LZG scanner kit,[1] and having previous experience in simplifying communications circuitry by eliminating ICs, I redesigned the scanner circuit with discrete components. This produced many unexpected side benefits.

The scanner built with discrete components (fig. 1) offers the following advantages, in addition to availability and cost of components.

1. Simpler type of readout.

2. Ease of maintenance – troubleshooting is easier with discrete components and spare parts are no problem (almost any silicon transistors provide proper operation).

3. Unit operates directly from a 10- to 15-volt power supply. IC version requires a 5-volt regulator.

4. Less current drain – about 8 times less than the IC version with regulator.

5. Smaller – PC board layout is one-quarter to one-third the size of the IC version. Actual size of board is 2x2-1/8 inches (51x54 mm). This is primarily due to the flexibility of discrete components in crossover-free, small area layout.

6. Flexibility of design – unit can be adapted easily for less than the full four channels without having to skip channels between the active ones. The unit is also adaptable to any desired scan rate with a simple component change.

The new design also takes into account several variable factors which you may run into when applying the scanner to your particular receiver. The unit can be used with either positive or negative logic from your squelch circuit. That is, scanning can be stopped with either a positive or ground signal. The input circuit is very sensitive; therefore, almost any voltage, from a few volts up, may be used to operate the scan-stop circuit. The unit normally puts out +5 volts to turn on (*enable*) the oscillator. However, simple additional inverter transistors used with the oscillator turn-on circuits allow the scanner to operate with oscillator circuits requiring a ground signal to *enable*.

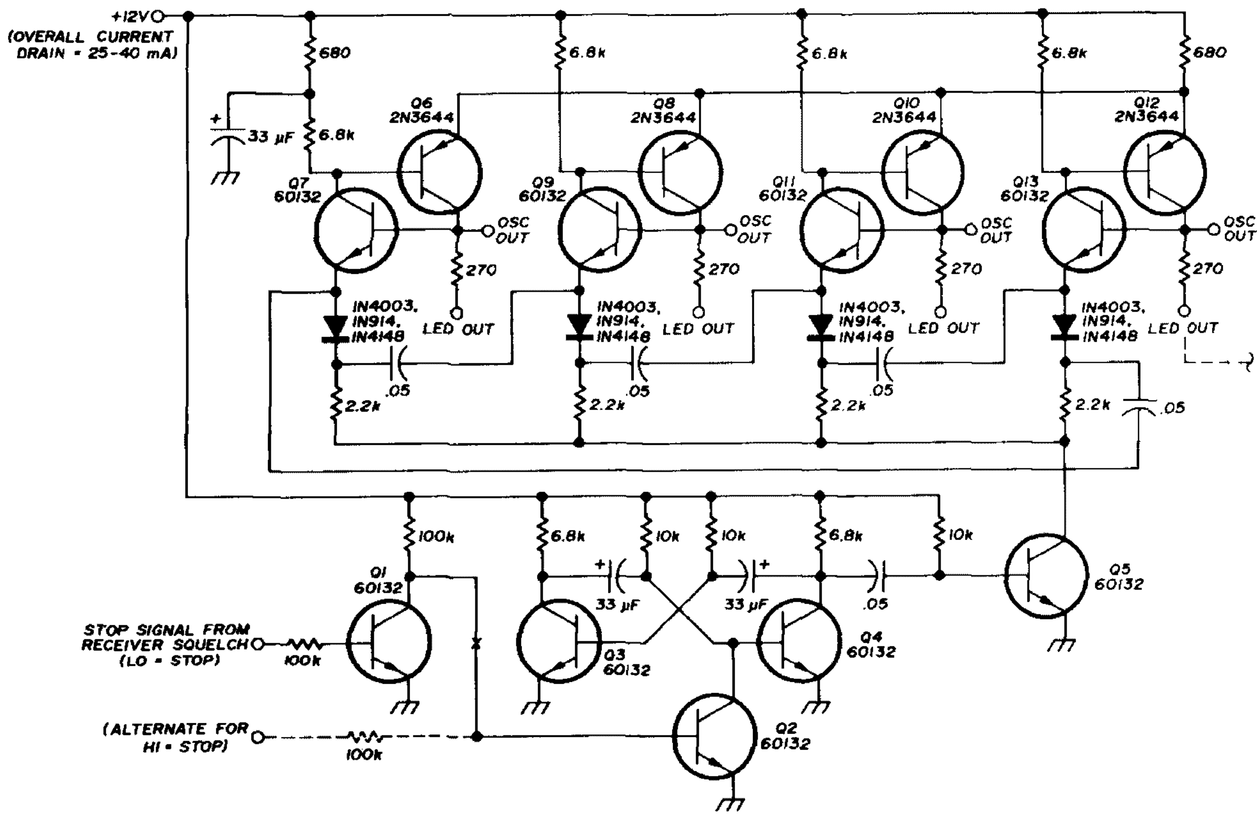


fig. 1. Schematic diagram of the improved fm receiver scanner. Additional channels may be added by connecting additional ring-counter stages, as discussed in the text.

typical test voltages, volts (ground reference)

Q1 collector	0.6/0
Q3/Q4 base	0/0.6
Q3/Q4 collector	0/12
Q5 base	0.6
Q5 collector	0.1
Q6/Q8/Q10/Q12 base	12/4
Q6/Q8/Q10/Q12 collector	0/5
Q7/Q9/Q11/Q13 emitter	0/3.8

theory of operation

Transistors Q3 and Q4 form an astable multivibrator, operating at a pulse repetition rate of approximately 10 pulses per second. Transistor Q2 turns *on* to stop the multivibrator when Q2's base is *high*. Q2 does this by shorting the base of Q4 to *ground*. Normally, squelch circuits have a *high* output available when closed and a *low* output when open. Therefore, inverter Q1 is used to provide the proper polarity signal to operate Q2. Transistor Q1 is tied to the collector of the squelch switch stage in the receiver through a 100k isolation resistor to turn on Q1 when the squelch is closed. When the squelch is open, Q1 is turned *off*, Q2 is turned *on*, and Q4 is locked *off*. This stops the sequencing action of the scanner. For those squelch circuits operating in the opposite sense, the alternate circuit connections shown on the schematic may be used. Transistor Q1 is deleted, and Q2's base is operated directly from the receiver.

Transistors Q6 to Q13 form a four-stage ring counter. Normally each stage is turned off by the base-emitter biasing on the pnp transistor. When power is first applied, the 33-μF capacitor in the biasing circuit for the base of Q6 ensures that the first stage will turn on by momentarily upsetting the normal bias. Subsequently, due to the return for ring counter stages being pulsed by Q5, the *high* output is passed from one stage to the next in an endless ring pattern. Normally Q5 is turned *on* hard to ground, and it is pulsed to a *high* condition for a short pulse duration each time a negative-going pulse is received from the multivibrator. When the pulsating action ceases due to operation of the squelch circuit, the ring counter stops stepping, holding the output of one ring counter stage *high* until the squelch circuit closes.

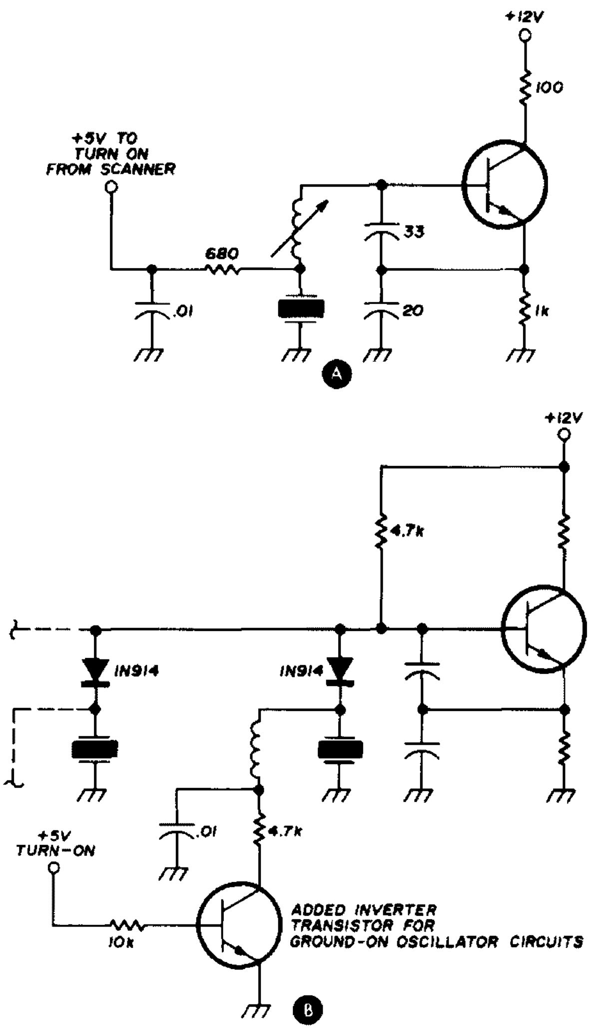


fig. 2. The receiver scanner shown in fig. 1 may be used with oscillators that require +5 volts for turn-on (A), or for oscillators requiring a ground for turn-on (B).

The collector of the pnp transistor in each stage provides a +5 volt output directly to the oscillator to be enabled and also through a 270-ohm current-limiting resistor to the LED readout corresponding to the enabled channel. It is important to note that the 270-ohm resistor in each stage must be connected to the LED for proper biasing to be maintained. If LEDs are not used, adjustments in the circuit or dummy resistor loads must be used to maintain the same current flow in other parts of each ring-counter stage.

The outputs of the scanner can be used to turn on oscillators in various ways, depending on the particular oscillator. Since this subject has been treated in several magazine articles,[1,2] only a few examples will be given. Refer to fig. 2 for examples of two common types of oscillators which may be encountered. The examples illustrate, in general, how oscillators may be enabled by the scanner.

The first circuit (fig. 2A) shows a typical +5 volt turn-on type of oscillator. This circuit is used in a vhf fm receiver described in a previous article.[3] The oscillator has no internal bias; therefore, it cannot oscillate by itself. The base of the transistor, connected through a coil to the crystal and to the *enable* line through a resistor and 0.01-μF bypass capacitor, is biased on by +5 volts from an external source. The resistor and bypass capacitor block the flow of rf on the control line. Application of approximately +5 volts from a scanner circuit to the control line will turn the desired oscillator on.

The second circuit (fig. 2B) shows a conventional oscillator circuit adapted to multichannel operation by using diode switching. This circuit was commonly used a few years ago to modify older tube-type equipment for multichannel operation. In this circuit the crystal is essentially turned on at appropriate times by grounding the corresponding control line. In this case each leg is turned *on* when desired by a *ground* from an added

inverter transistor which turns *on* when +5 volts is applied to its base from the scanner through an isolation resistor.

It can be seen from these examples and the referenced articles that almost any fm transceiver or receiver can be modified to provide scanning operation. Application of one or more of the ideas illustrated can be used to convert almost any oscillator circuit.

scanner circuit modifications

There are at least four possible modifications which may be of interest to other system requirements, if necessary.

1. Less than four-channel scan. This may easily be accomplished by reconnecting the 0.05-μF capacitor at the bottom of the last ring-counter stage to be used. It should be rewired to the first stage instead of the next stage. This effectively shortens the ring. The Q6/Q7 stage must always be used since the bias circuit for the base of Q6 includes a capacitor and resistor which ensure that the ring counter operates when power is first applied. On the PC board shown in **fig. 3** this is best done by reconnecting the last 0.05-μF capacitor to the line along the bottom of the board running back to the first stage.

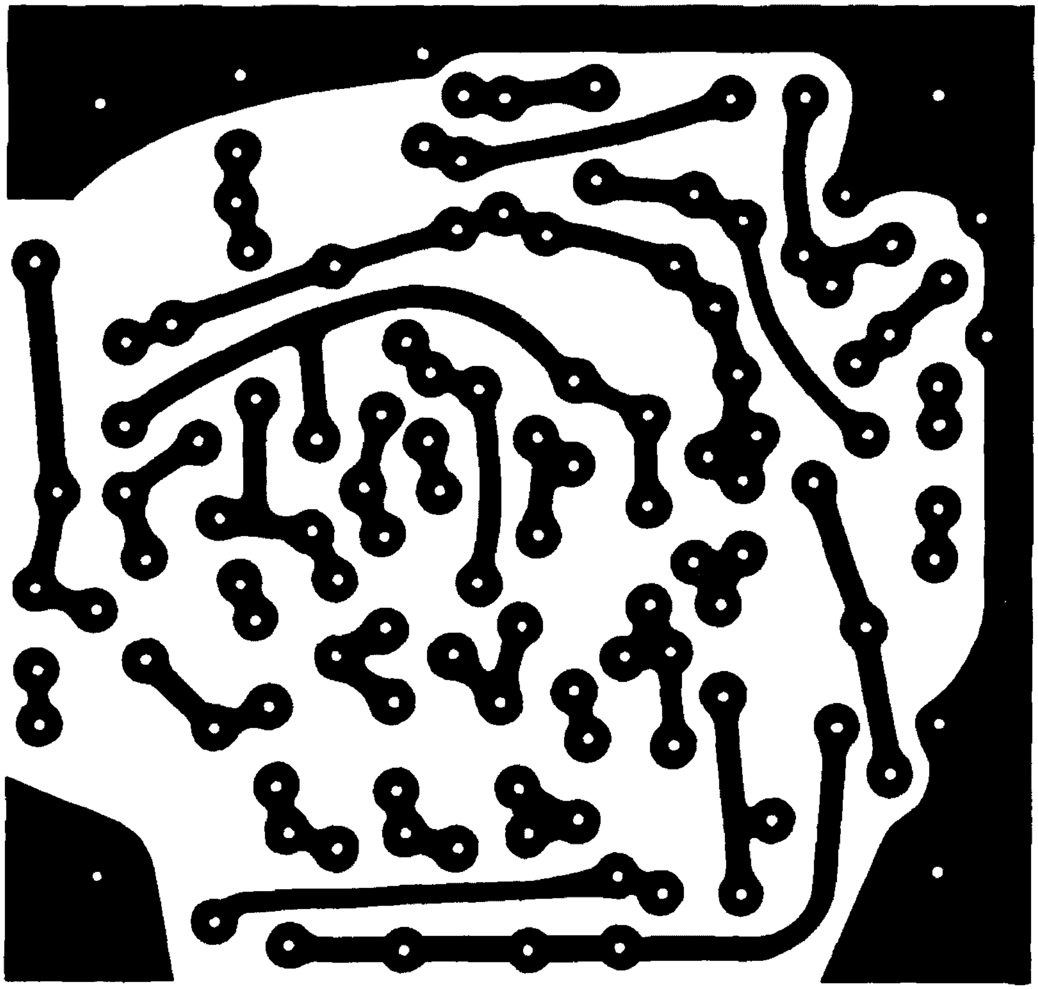


fig. 3. Printed-circuit layout for the improved fm receiver scanner. Component layout is shown in fig. 4.

2. Changing scan speed. The speed was set up to be the optimum compromise between proper squelch triggering and ability to catch short transmissions. If the approximate 10 channel/second rate requires change for some reason, different values for the two 33-μF capacitors in the multivibrator will change the oscillation rate of that stage.

3. Increasing number of channels. If you can stand the confusion of scanning more than four channels, more than one board may be used. Link the ring counter stages on the two (or more) boards, and use only the triggering and timing circuits of the first board.

4. Programming channels. Switching schemes can be used between the scanner pads darkened in on the layout drawing, fig. 4, and the top leads are looped back down to the pads which are circled. Polarity should be observed on the diodes and electrolytic capacitors. Install the 2N3644 transistors first to avoid any mixup later with the 60132 (general purpose npn) transistors.

Pads which are circled and have an X in them are inputs and outputs and will

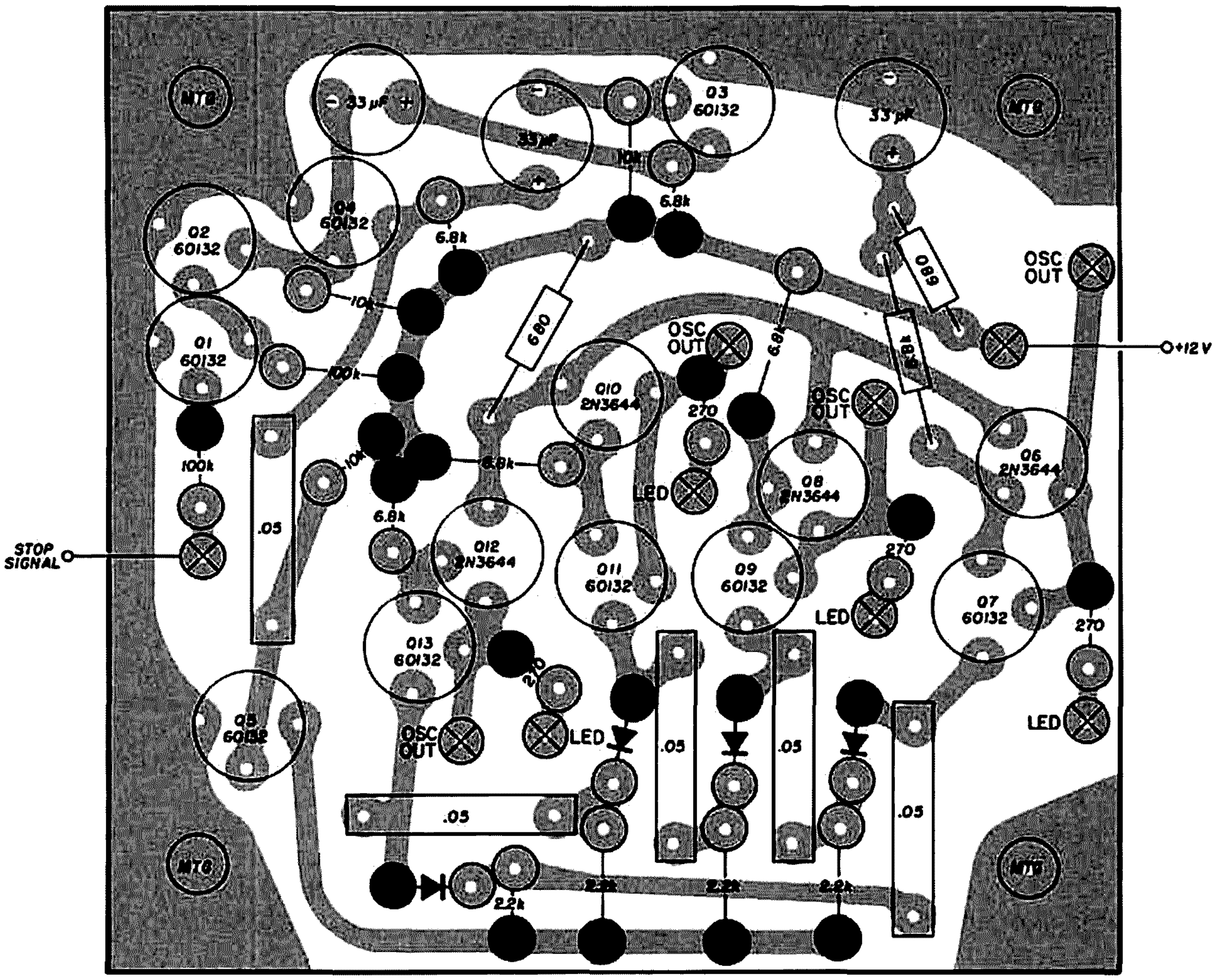


fig. 4. Component layout for the improved fm receiver scanner. Dark circles indicate bottom of resistors (or diodes) installed in vertical position; the other lead is looped over to the open circle.

outputs and the various oscillators to disable one or more channels at certain times or to select particular channels to be monitored out of a bank of the available channels.

construction

Assembly is straightforward. If the etched circuit board shown in fig. 3 is used, most of the resistors and all of the diodes are installed standing up on the accept wire such as number-22 hookup wire, for connections to the outside world. The +12-volt terminal should be connected to a source of filtered +10 to +15 Vdc. The input terminal should be connected to the collector of the receiver's squelch switch stage. If the alternate input is used, Q1 and the 100k resistor from +12 volts to the base of Q2 can be eliminated and the 100k resistor from the input terminal can be connected

directly to the base of Q2. The oscillator outputs should be connected to individual oscillator stages, and the LED outputs should be connected to the anodes of the LEDs.

The cathodes of the LEDs should be connected to ground. Using the type of LED supplied in the kit,* the orange dot on the base indicates the cathode terminal. Be sure you observe polarity or the LED won't illuminate. The LEDs supplied with the kit may be mounted by inserting them in 13/64-inch (5-mm) holes drilled in the front panel of the radio. A spot or two of epoxy cement on the base of each LED behind the panel should hold them in place. Alternate LEDs sometimes have one short lead to indicate the cathode. Likewise, alternate 1N4148 diodes used in ring counters have their cathodes identified by a shorter lead.

If you wish to use both the scanner and a regular channel switch, this may be done by first connecting the oscillators to operate with the scanner and then wiring up a rotary channel switch to supply +5 volts to select oscillators when the scanner is turned off. A five-position switch can be used for this purpose with the fifth position turning on the scanner. In positions 1 through 4, the scanner is turned off, and +5 volts is connected through the switch to the oscillators corresponding to the selected channel positions on the switch.

references

1. Bob Reifsteck, K2ZLG, "VHF Receiver Scanner," *ham radio*, February, 1973, page 22.
2. George Allen, W2FPP, "VHF FM Channel Scanner," *ham radio*, August, 1971, page 29.
3. Jerry Vogt, WA2GCF, "VHF FM Receiver," *ham radio*, November, 1972, page 6.

ham radio

*In conjunction with this article a kit is available, complete with undrilled G-10 PC board, all components, the LED indicators and an instruction manual. Price of the kit is $10, including domestic parcel post. If desired, add 50¢ for a number-66 drill bit or 40¢ for air-mail delivery. To order, or to obtain information on other fm kits, write to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

how to measure peak envelope power

Don't depend on your meters for *accurate* measurement of PEP input power — use the technique discussed here

Carl Drumeller, W5JJ, 5824 NW 58 Street, Warr Acres, Oklahoma 73122

It's unfortunate that the FCC requires amateur radio stations to compute power input for logging purposes. This is simple for fm or CW or even the almost-obsolete "conventional a-m" (6A3). For the user of single sideband (3A3J), however, the undertaking is far from simple. The purpose of this article is to identify those aspects that contribute to the difficulty of measurement and to present methods whereby the measurement may be made with a reasonable degree of accuracy.

Because the input power to an ssb transmitter continually varies when transmitting voice signals, custom and law stipulates that peak input power must be that which is measured. How do you catch that fleeting, almost ethereal peak of power induced by the upper excursion of a complex waveform? No meter can follow it without unpredictable lag or overshoot. Even if it could, what eye could register a needle flick of less than a thousandth of a second? Being knowledgeable of the instantaneous plate current, assuming you're using a vacuum tube in the metered stage, is, by itself, a formidable undertaking. But power computation also requires knowledge of instantaneous plate voltage. Unless the transmitter's power supply has superb dynamic regulation, far better than you have any right to assume, you're faced with another (but not quite so formidable) task of super-quick information gathering.

We've coasted along for years with the possibly-right, probably-wrong assumption that the peak needle excursion of a plate current meter equals one-half of the actual peak current. The FCC gives a quasi blessing to this assumption by accepting it, but only if the meter is that supplied by the original equipment manufacturer. It's highly doubtful that any FCC engineer has illusions regarding the accuracy of this method of ascertaining peak plate current. It's accepted only because not accepting it would rule off nearly every ssb transmitter in the Amateur Radio Service.

Unless the input power is within 10% of the maximum permitted, the FCC will also accept the nominal *rated* voltage of the plate power supply.

Now, let's review what we have for making our educated guess as to the peak

input power of a voice-modulated ssb transmitter:

1. A plate meter indication that is dependent upon the dampening characteristics of the meter as well as upon the voice characteristics of the speaker. These voice characteristics vary not only from person to person, but also as to type of speech used by the individual. If you talk close to the microphone, with the gain turned up, using a monotone normal to confidential conversation, the meter will give one type of deflection. On the other hand, if you're back from the microphone, talking in a quick, excited tone, the deflection will be quite different for a given *peak* excursion.

2. The plate voltage may be measured, or it may be taken from the manufacturer's specifications. If measured, it may have been under *no load* condition. If under loaded conditions, the fall-off under syllabic current loads may possibly be gradual enough to be read on the voltmeter. Unless a huge filter capacitor is used, it's probable that voice-peaks markedly load down the voltage.

The probabilities of truly accurate measurement of peak input power are about the same as those of rolling a seven ten times running at a Las Vagas gambling table! How, then, can you make an accurate measurement?

Assuming, for a start, that you have a child-like trust in the utter trueness of the stated plate voltage, I'll delve into the matter of plate current measurement. Let's hope your transmitter permits an easy access to its negative high-voltage lead so that this lead may be broken for the insertion of a small resistor. Otherwise you'll need, in addition to the normal equipment, an isolation transformer (good for quite high voltage), a well-insulated oscilloscope cart and the type of constitution that permits playing Russian roulette!

But let's say your transmitter is metered in the negative lead, as are most modern designs. You break the lead at a point that does not present any unintentional shunt paths, insert a resistor of a few ohms, connect an oscilloscope across this resistor, and then calibrate the scope in terms of current. It's best to use a scope with a long-persistence CRT. This lets you take a deliberate view of a trace that flicks across the face in less than a thousandth of a second. Now, as you speak into the microphone, you'll be able to see the maximum current drawn, reading it off the calibrated oscilloscope graticule.

For a better idea of your peak input power (if you can scrape up another scope), you can build a resistive voltage divider across the output of your power supply, tapping off a little potential for your scope input. Here, again, you have a simple calibration job to do. With two measuring devices which are not burdened by inertia and yet hold a reading (peak for current, dip for voltage) long enough to permit accurate observation, you're all set to read peak input power. (Provided that you're using a common-cathode triode operating in class A or class AB1.)

Have you noticed the way the FCC is currently grading its operator examinations? Power input to a vacuum tube no longer is measured by just its plate power input. You must also consider the power fed into the screen grid (if used), the suppressor grid (if used) and the control grid! That last is rf power. The others are dc, of course.

Most high-power rf amplifiers use common-grid triode vacuum tubes (also called grounded-grid). If yours does, coming up with the total power input is comparatively easy. All you need to do is to first measure the vswr on the line between the exciter and final. Make it unity. Then measure the instantaneous rf voltage on it with another calibrated oscilloscope. Compute the rf power by E^2/R, where R is the cable impedance. Add this power to the measured dc input power, and you should have fully satisfied all the requirements for ascertaining the peak input power of your ssb transmitter!

ham radio

how to predict harmonic output

Determining drive points for optimum harmonic generation

Fin K.L. Utne, Fjellvn. 7, 1800 Askim, Norway

Why are certain oscillator circuits better suited for frequency doubling than others? Why are other circuits more suitable for frequency tripling? Questions like these have bothered many amateurs and experimenters. It is the aim of this article to shed some light on the subject of the harmonic content of some common non-sinusoidal waveforms.

A signal of pure sine waves contains only its fundamental frequency. Any departure from this sine-wave pattern, no matter how small, is due to the presence of additional frequencies that are multiples (harmonics) of the fundamental frequency. If this non-sinusoidal wave can be given an exact mathematical description, the amplitude and phase of each harmonic frequency it contains may be calculated.

analysis

For example, assume that we have a series of waves, all alike, each looking like that pictured in fig. 1. This is a sine wave

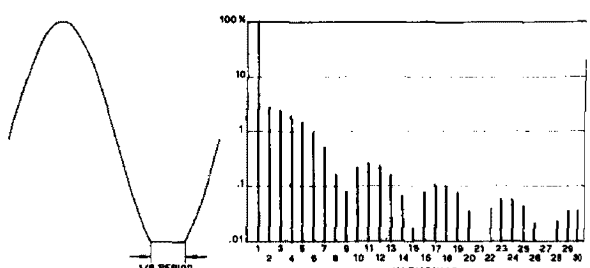


fig. 1. Sine wave with a small amount of negative peak clipping. Graph shows relative amplitudes of the resulting harmonic products.

with the tip of the lower loop sliced off. Such a waveform might be produced when an amplifier is driven slightly beyond the cutoff point.

A Fourier analysis will show that this deformation of the sine wave gives rise to an endless number of harmonics, with amplitudes that tend to decrease with growing frequency. However, this tendency is not uniform.

The chart in **fig. 1** illustrates the relative amplitude of each harmonic frequency up to the 30th. The scale is logarithmic, and the columns show the amplitudes of the harmonics as a percent of the original waveform's amplitude.

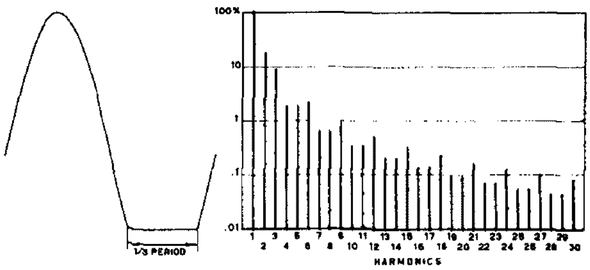


fig. 2. Harmonics generated when the lower portion of a sine wave is sliced off for one-third of the period. Maximum harmonic output occurs at harmonics which are third multiples (3, 6, 9, etc.) of the fundamental frequency.

As mentioned before, the general trend is for the amplitude to decrease with increasing harmonic number, but this tendency is not uniform. A curve joining the tops of the relative amplitude columns makes a kind of wave pattern of its own. with minimum values at the 9th, 15th, 21st, and 27th harmonics.

Fig. 2 shows the situation when a still larger portion of the lower half of a sine wave is sliced off. In this case it would seem that an amplifier is driven into cutoff for a third of the duration of a period. In this case harmonic content is maximum for every harmonic divisible by three. Peculiarly, the two intervening harmonics have amplitudes of equal magnitude.

Another point worth noting is that the fundamental frequency (or the first harmonic) has an amplitude about 7% larger

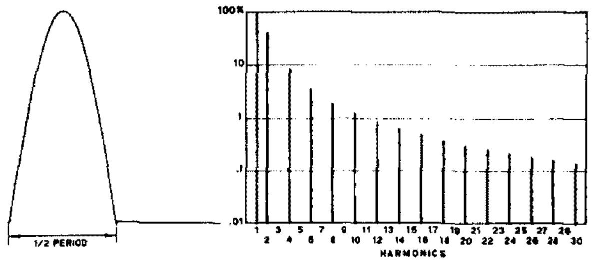


fig. 3. Sine wave with entire negative portion clipped off (rectified half-wave). Note that odd-order harmonics are eliminated entirely.

than the original deformed sine curve. What a temptation to drive an amplifier just a trifle over saturation, to get that little extra gain at the fundamental frequency! Actually, for this series of idealized waveforms the fundamental frequency increases to a maximum of 107.305% when the portion of the sliced-off sine wave represents 31.9% of the period.

half sine waves

Fig. 3 shows the case where the lower excursion of the sine wave has been clipped off entirely—a rectified half wave. Note that every harmonic with an uneven number has disappeared. The fundamental frequency has an amplitude equal to the actual amplitude of the half wave.

Fig. 4 shows a sine wave sliced off even more, so that only a third of the period is left. The envelope over the harmonic columns shows a pattern much like **fig. 2**, but the fundamental frequency has an amplitude of only 78% of the curve's amplitude. The higher harmonics, though, have larger amplitudes than those of **fig. 2**.

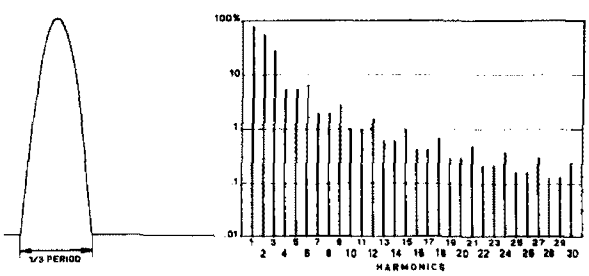


fig. 4. Raising the clipping point above the zero-crossing point sharply increases harmonic output. Note resemblence of this spectrum to that of fig. 2.

In fig. 5 the sine wave is sliced off so much that only one-sixth of the period remains. The fundamental frequency has an amplitude of only 43% of the curve but the harmonics are quite prominent. As in fig. 1, the envelope over the harmonic amplitude columns shows a wave pattern with minima at the 9th, 15th, 21st and 27th harmonic. For more precise harmonic amplitude values for fig. 1 through fig. 5, see table 1.

what this means

Now we can return to the questions asked at the beginning of this article. Why are certain circuits better suited for frequency doubling than others? Looking at table 1, you can see that the second *harmonic in the column* under 2/3 is larger than all the others. This means that slicing off two-thirds of the sine wave would give better second harmonic performance than any of the other clipped waveforms pictured. Actually, for the highest second harmonic content, 66.68% of the sine wave must be clipped off. This represents a conduction angle of 119.9° and provides a second-harmonic amplitude of 55.133% of the basic curve's amplitude.

table 2. Sine waveform clipping required to optimize different harmonics.

harmonic number	clipping required	conduction angle	amplitude (maximum)
1	31.9%	245.1°	107.305%
2	66.7%	119.9°	55.133%
3	77.9%	79.7°	36.908%
4	83.4%	59.8°	27.719%
5	86.7%	47.7°	22.190%
6	89.0%	39.8°	18.498%
7	90.5%	34.1°	15.859%
8	91.7%	29.9°	13.878%
9	92.6%	26.5°	12.337%
10	93.4%	23.9°	11.104%
11	94.0%	21.7°	10.095%
12	94.5%	19.9°	9.254%

table 1. Harmonic amplitudes as a percentage of the basic sinusoidal waveform for various amounts of clipping.

	portion clipped off bottom of sine waveform				
harmonic number	1/6 (fig. 1)	1/3 (fig. 2)	1/2 (fig. 3)	2/3 (fig. 4)	5/6 (fig. 5)
1	104.089	107.267	100.000	78.200	43.045
2	2.843	18.378	42.441	55.133	39.598
3	2.462	9.189	0.000	27.566	34.293
4	1.990	1.838	8.488	5.513	34.293
5	1.477	1.838	0.000	5.513	27.719
6	0.975	2.100	3.638	6.301	20.576
7	0.528	0.656	0.000	1.969	13.577
8	0.169	0.656	2.021	1.969	7.349
9	0.082	0.919	0.000	2.757	2.357
10	0.224	0.334	1.286	1.002	1.143
11	0.269	0.334	0.000	1.002	3.120
12	0.239	0.514	0.890	1.542	3.741
13	0.162	0.202	0.000	0.606	3.323
14	0.069	0.202	0.653	0.606	2.261
15	0.018	0.328	0.000	0.985	0.957
16	0.079	0.135	0.499	0.405	0.245
17	0.109	0.135	0.000	0.405	1.106
18	0.106	0.228	0.394	0.683	1.513
19	0.078	0.097	0.000	0.290	1.471
20	0.036	0.097	0.319	0.290	0.506
21	0.006	0.167	0.000	0.501	0.089
22	0.040	0.073	0.264	0.218	0.559
23	0.058	0.073	0.000	0.218	0.813
24	0.059	0.128	0.221	0.384	0.826
25	0.045	0.057	0.000	0.170	0.633
26	0.022	0.057	0.189	0.170	0.311
27	0.003	0.101	0.000	0.303	0.042
28	0.024	0.045	0.163	0.136	0.336
29	0.036	0.045	0.000	0.136	0.507
30	0.038	0.082	0.142	0.245	0.529

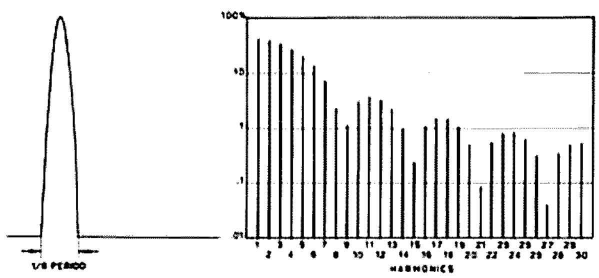


fig. 5. Severely clipping a sine wave so that only the positive peak remains results in lower-order harmonic output very nearly as strong as the fundamental.

In table 2, the optimum slice-fractions (conduction angles) are given for harmonics up to the 12th, with their corresponding highest possible amplitude for this set of non-sinusoidal curves.

summary

An oscillator producing an output curve in the form of a horizontally amputated sine wave will provide good second harmonic content if the horizontal part of the curve is approximately two-thirds of the total period. If frequency tripling is wished, the horizontal part of the curve (that is, the duration of the cutoff) should be in the neighborhood of three-quarters of the total period. Conversely, if a specific harmonic must be minimized the data of table 1 can be used to determine the best operating point.

APPENDIX

A Fourier-series of the form $f(X) = \frac{1}{2}a_o + a_1\cos X + a_2\cos(2X) + \ldots$ for a sine wave clipped off by a fraction $2V$ of the total period 2π will have the following coefficients:

$$a_o = [2/(1+\cos V)]\,[2/\pi]\int_V^\pi (\cos V - \cos X)\,dX$$

$$= [4/\pi(1+\cos V)]\,[(\pi - V)\cos V + \sin V]$$

$$a_1 = [2/(1+\cos V)]\,[2/\pi]\int_V^\pi (\cos V - \cos X)\cos X \cdot dX$$

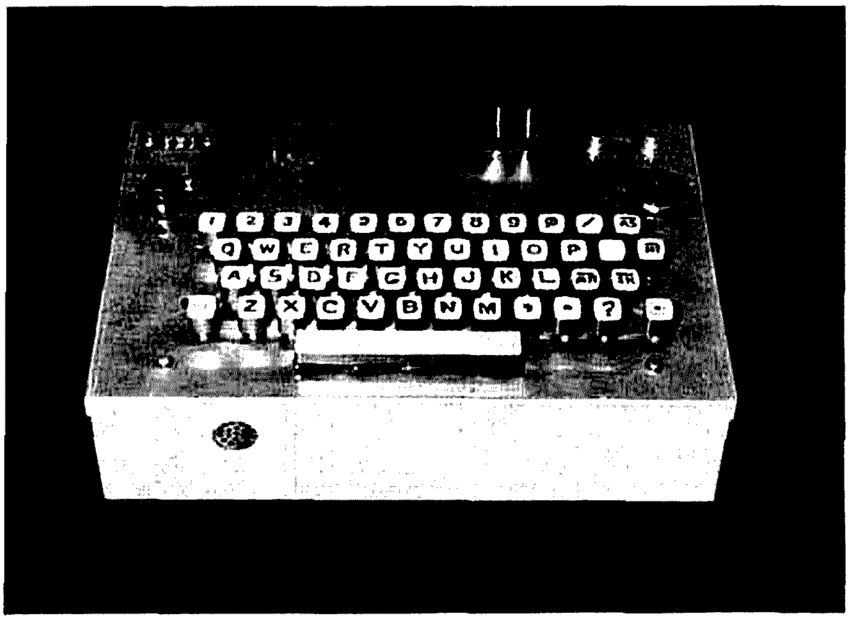
$$= [2/\pi(1+\cos V)]\,[V - \pi - \sin V \cdot \cos V]$$

$$a_n = [2/(1+\cos V)]\,[2/\pi]\int_V^\pi (\cos V - \cos X)\cos(nX) \cdot dX$$

$$= [2/n(1+\cos V)]\left[\frac{\sin(n+1)V}{n+1} + \frac{\sin(n-1)V}{n-1} - \frac{2\sin(nV)\cos V}{n}\right]$$

a_o represents the value of the direct current present, and has no significance in this discussion. The absolute values of a_1, a_2, and so on, give the relative amplitudes of the corresponding harmonics.

ham radio



the code mill

A Morse keyboard that provides perfect character generation as well as perfect letter and word spacing, regardless of the operator's keyboard skill

H.E. Thomas, W6CAB, 33020 Almond Street, Lake Elsinore, California 92330

Almost without exception, any person who uses Morse code can attain highest copying speed only when the code is sent perfectly. Because of this, and also in many cases as a matter of pride, most amateurs have tried to improve their sending technique. Many different kinds of equipment have been devised which can assist in this, but none of these, reasonably available to the average amateur, can produce perfect code, despite claims to the contrary.

Perfect International Morse code consists of five parts: the dot, which is one time unit or baud; the dash, which is three bauds; the space between each element of a Morse character, which is one baud; the space between letters, which is three bauds; and the space between words, which is seven bauds. Each and every one of these five parts of International Morse must be produced with machine-like precision – not a single one can be produced by human estimation if *perfect* code is to be the result.

The typical bug can make dots quite well but the other parts of the code are up to the operator. Some types of electronic keyers can produce dots, dashes and the spaces between them in proper form, but the operator is responsible for

the letter and the word spaces. The latest device, which goes a step beyond the electronic keyer, is what I call a *code mill*. In radio parlance a typewriter is a *mill*, so what better name for keyboard keyers?

These code mills create perfect dots, perfect dashes, perfect spacing between elements of the characters and terminate each character with a perfect letter space. Whether this letter space remains perfect depends on the availability of the succeeding letter at precisely the right time. This in turn depends on whether the operator is successful in punching the key of the succeeding letter during what might be called its *launch window*. This launch window becomes quite narrow as the code speed increases and is almost intolerably narrow for any character which may follow the letter E (the launch window in this case is less than a tenth of a second at 30 wpm).

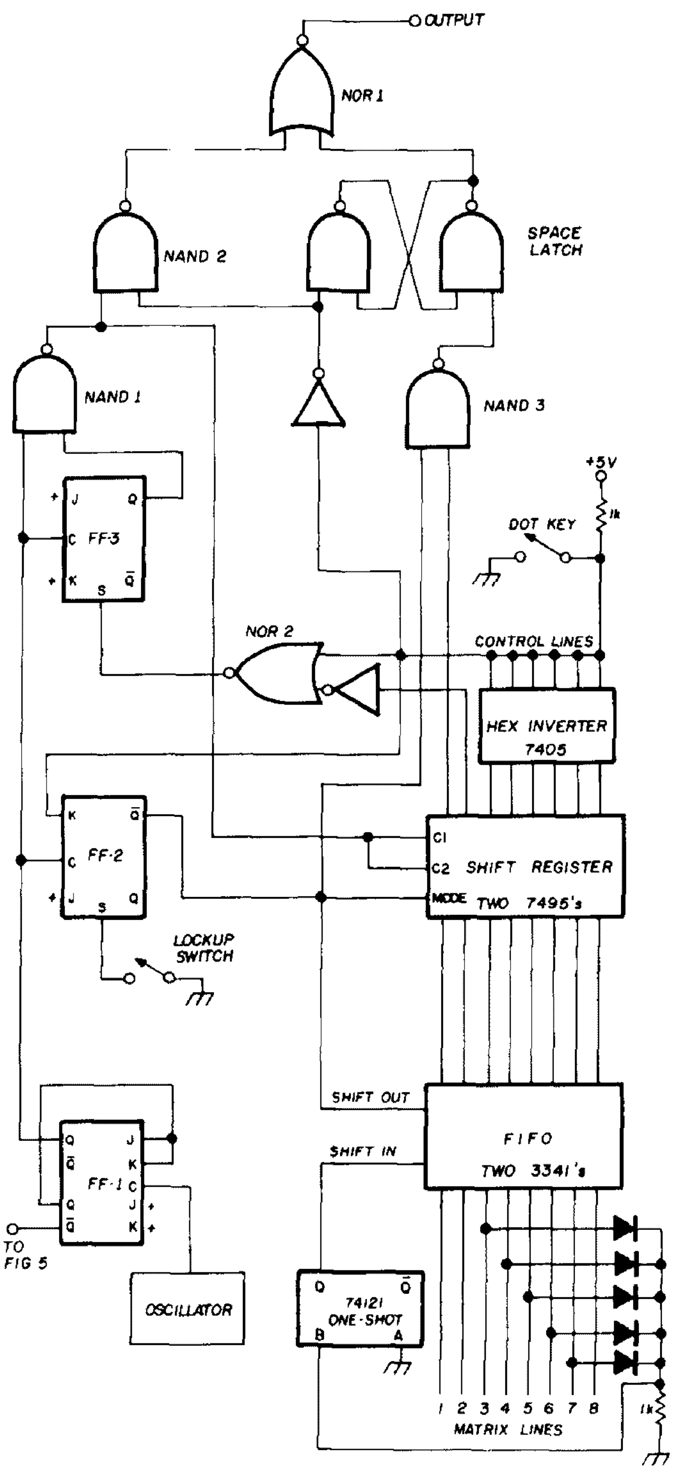


fig. 1 Block diagram of the code mill which uses first-in first-out (FIFO) serial memory to generate perfect Morse code characters and spacing, regardless of the operator's keyboard skill.

Some keyboard keyers provide a one-character storage latch at the input of the circuitry but even this partial solution requires the operator to anticipate various letter combinations. The varying length of the different Morse characters is not compatible with the smooth rhythm of good typing, particularly at the higher speeds. And, of course, the usual code mill does nothing whatsoever about word spacing. Listening on the air will give ample evidence of the inability of such devices to produce perfect code.

the solution

The obvious answer to the problem is to design a code mill which will automatically produce all five parts of the Morse code. Since this includes word spaces, a space bar for the keyboard is required, as in teletype equipment. Furthermore, to maintain exact letter spacing, the availability of successive characters must also be automatic. This means that the operator and his sometimes erratic operation of the keyboard must be isolated from the code transmitting circuits.

The electronic equivalent of a vending machine can best meet this requirement. The keyboard operator dumps the merchandise in at the top in the proper sequence and at whatever rhythm suits, including the "hunt and peck" system. The code transmitting circuit puts the "coin in the slot" and extracts the merchandise in the same sequence but at precisely the correct instant.

the circuit

This and the following paragraphs describe a code mill with an electronic vending machine having a storage capacity of 64 Morse characters. This code mill produces perfect code at any speed between 7 and 70 wpm as long as the operator keeps the storage partially filled and does not exceed its storage capacity at any time.

The block diagram of fig. 1 shows the basic arrangement which uses two Fairchild 3341 mos ICs for storage. Each 3341 is a 64 by 4-bit FIFO (first-in first-out) serial memory, and is available in a standard ceramic 16-pin dual-inline package. A detailed description of circuit operation will not be given here since the circuit, excluding the FIFOs, is a modernized version of a circuit which was described in fine detail in a previous article.[1] However, a brief description follows.

The oscillator is a variable-speed pulse generator which runs continuously. It triggers dual flip-flop 1 whose second Q output gives the required square wave. This square wave will produce transmitted dots through gates NAND 1, and NAND 2 and NOR 1 if none of these gates is inhibited. When no character is being processed through the shift register, the control line will be *high* and this inhibits NAND 2 via an inverter. Also, with the control line *high*, the K input of flip-flop 2 is *high* and this flip-flop is then triggered back and forth by the output of flip-flop 1. Whenever the $\bar{Q}$ output of flip-flop 2 is *high*, the FIFOs and the shift registers are ready to process Morse characters for transmission.

At this point it should be mentioned that the mos chips used in this circuit require a bit higher than normal *high* level. For this reason flip-flop 2 is one-half of a Signetics SP321A dual J-K flip-flop. Flip-flop 3 is the other half. Despite the fact that the Fairchild 3341 has internal pull-up circuits to make them compatible with TTL logic, the 3.8 volts minimum *high* of the SP321A looks a lot better than the 2.4 volts minimum *high* for the TTL flip-flops. A 7473 TTL could probably be used to replace the SP321A, but it was not tried.

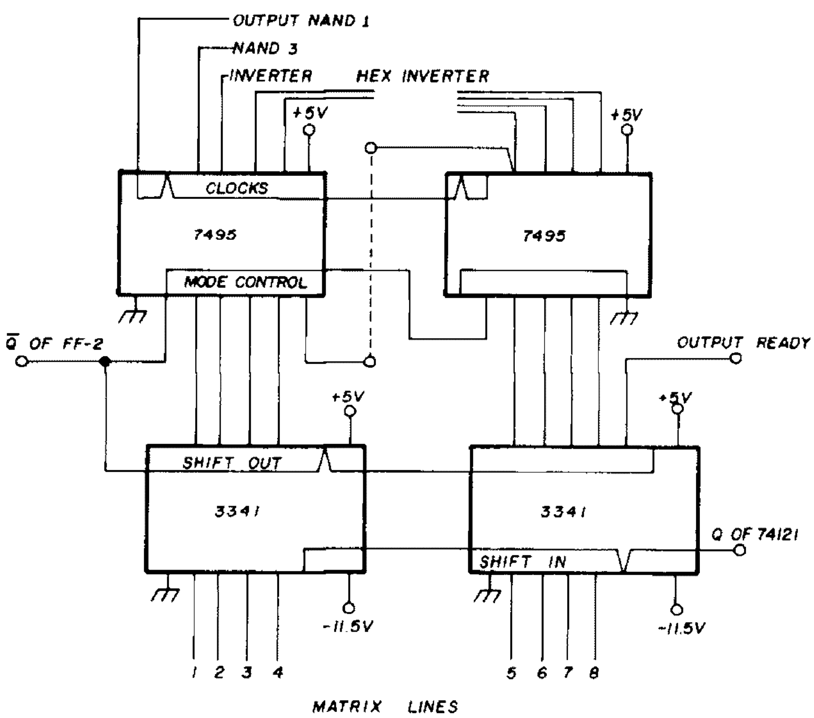
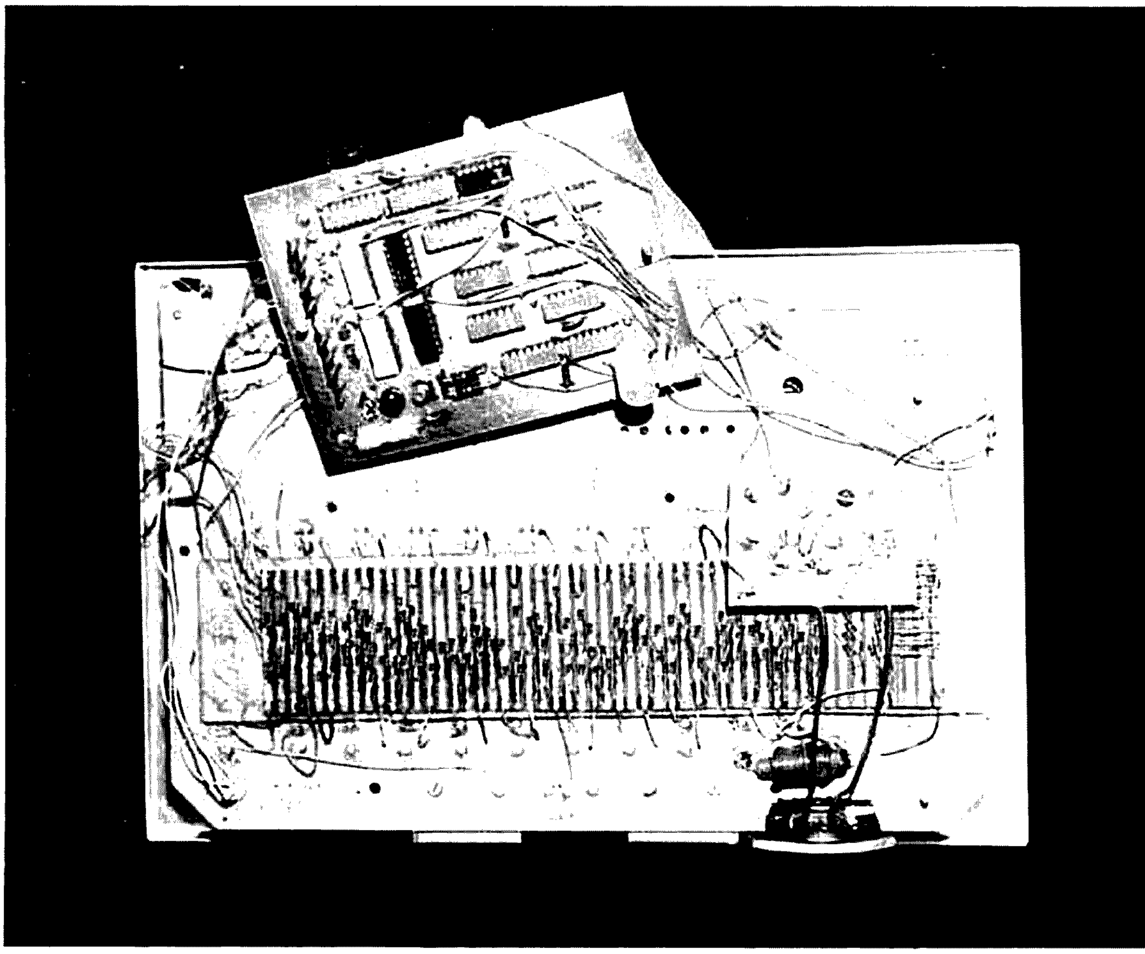


fig. 2. Bottom view, or foil side, of the ICs used in the first-in first-out serial memory and shift register.

The *set* input of flip-flop 2 may be grounded with a switch near the keyboard. This blocks the output of the mill while the FIFOs are being filled. With the lockup switch blocking the mill, calls may be typed up while still listening to the transmission from the other station or stations in a two-way or round-table QSO. It is also nice to use when setting up for a repetitive transmission using the recirculating shift registers to be described later.

When any key on the keyboard other than the space bar is depressed, the inputs to the FIFOs are actuated by the 74121

one-shot multivibrator, the character enters the FIFOs and almost instantly appears at the outputs. By the action of flip-flop 2 and the output of NAND 1 on the FIFOs' *shift out* controls and the shift registers' mode controls and clocks, the character enters the shift registers and eventually appears at the inputs of the hex inverter. This drops the control line output of the shift registers, the control line goes *low* because of the E, enabling NAND 2 and the left-hand gate of the space latch. The *high* on line 1, via NAND gate 3, sets the latch, which in turn inhibits NOR gate 1 and prevents transmission of the E. Note that NAND 3 is inhibited by the output of flip-flop 2 except during the loading of the shift



Construction of the code mill showing the matrix board mounted on the back of the keyboard switches. Oscillator is on small board to the right. Main circuit board, including memory, is at top.

to a *low* which enables NAND 2 (via the inverter) and the character is transmitted. NOR gate 2 and flip-flop 3 serve with NAND 1 to bridge the gap between two successive dots to produce a dash when required.

word space

When the space bar is depressed the letter E is formed and matrix line 1 is also activated (raised to a *high*). When the E and the *high* on line 1 appear at the registers. This prevents line 1 from setting the space latch when it goes *high* due to the shifting of *highs* along the shift registers. At the completion of every character the control line goes *high*, resetting the space latch through an inverter.

The letter E is used for the space character because it's one-baud length plus the three-baud automatic letter space adds to the previous three-baud letter space to become a seven-baud word

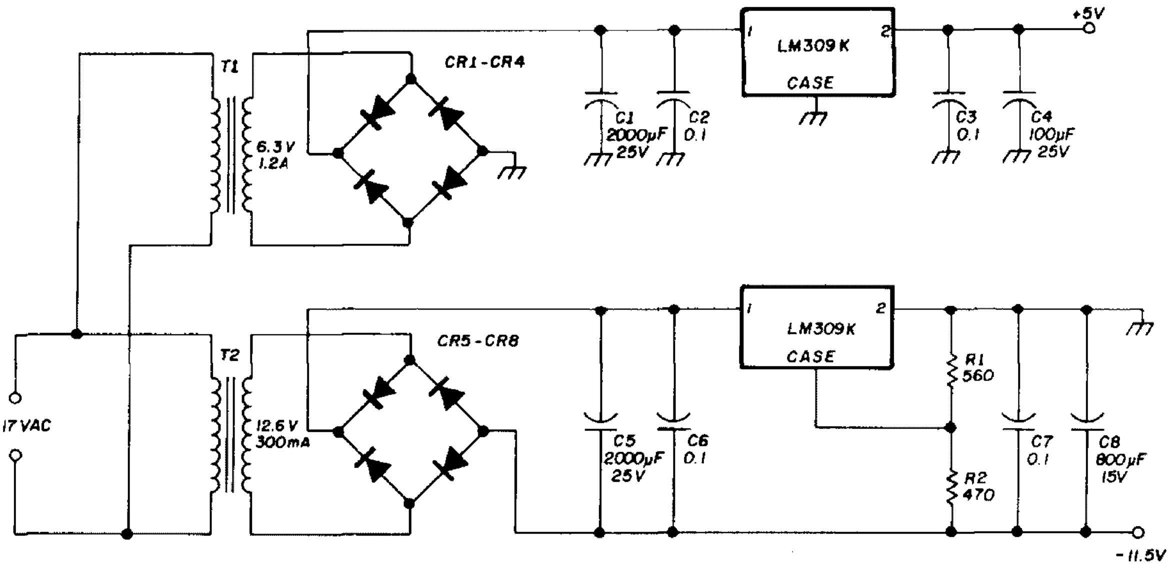


fig. 3. Regulated power supply for the code mill uses LM309K voltage-regulator ICs.

space. The letters I, S, H or the numeral 5 may be used if wider word spaces are desired. A switch is provided for this purpose, although it has seldom been used except while teaching code to beginners.

Fig. 2 shows the bottom view, or foil side, of the ICs comprising the FIFO and the register. Fig. 3 shows the circuit for the power supply. Both the +5 volt and the −12 volt supplies are regulated by LM309K voltage regulators. The values of the resistors in the −12 volt supply were chosen to give an output of −11.5 volts.

last-character elimination

Fig. 4 shows a circuit which was added some time after the code mill was built. Without touching the keyboard, this circuit will trigger the FIFO inputs at the conclusion of a character (control line goes *high*) and when the *output ready* terminal of one of the 3341s indicates that the FIFO is empty. Unfortunately, the 3341 will never completely empty itself. The last character will remain at the output of the FIFO until another character enters and "falls through" to the output. This causes continuous repetition of the last character. The operator must either learn to punch the space bar as a last character or a circuit must be designed to prevent this repetition. By triggering the input without punching a key, a complete blank is entered and passes through, and in turn takes the place of the offending character.

Since the Fairchild 3341 memories require +5 volts and a −12 volts in addition to the usual ground connection, it was a simple matter to add the recirculating shift registers shown in fig. 5. The reason flip-flop 1 is a dual unit is now apparent: it permits triggering the recircu-

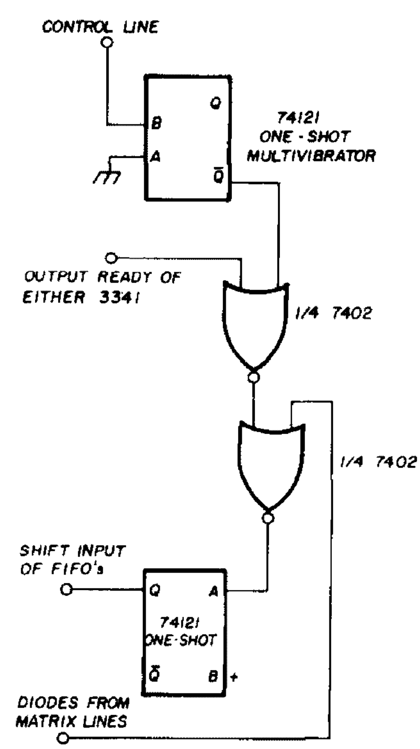


fig. 4. Circuit for last-character elimination.

lating shift registers with the output from the first $\bar{Q}$ which is equivalent to a trigger for every baud in the mill output.

The inverters at the input and output of the 2522 chips are used as buffers. The two 2522 chips will store every transmission from the code mill up to a limit of 528 bauds. When the spdt switch is thrown, the information stored in the 2522s will repeat over and over, and at any speed. Although not a necessary part of the code mill, this circuit is handy for preliminary callups prior to broadcasts, directional CQs or whatever, while you sip your coffee.

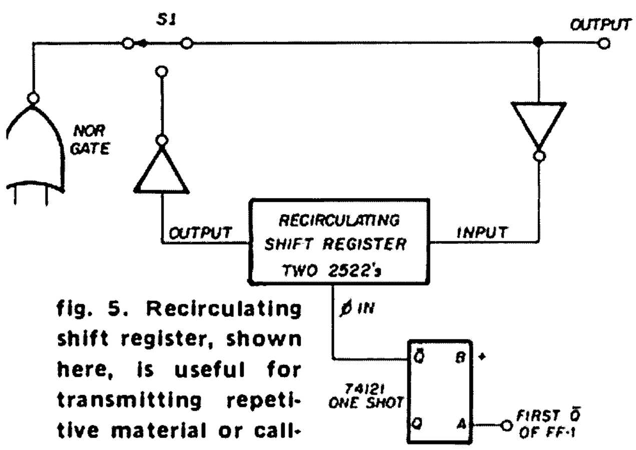


fig. 5. Recirculating shift register, shown here, is useful for transmitting repetitive material or calling CQ.

The code mill described in this article was built for just over one hundred dollars, and all chips were purchased at full list price. My keyboard itself is a surplus unit and similar types are available at several outlets. The price was not too high considering the very great pleasure and satisfaction derived from building the unit and operating it on the air. Without mentioning the code mill, almost every QSO elicits favorable comments on the keying. The code mill is just great for code practice, of course, but probably the best feature of all is the ability to send absolutely perfect code at fairly high speeds, hour after hour if need be, and with no great effort.

reference

1. J.A. Bryant, W4UX, "Touchcoder II," *QST*, July, 1969, page 12.

ham radio

automatic telephone controller for your repeater

A simple, reliable and foolproof control system for remote control of your repeater

Robert C. Heptig, KØPHF, Robert D. Shriner, WAØUZO

The problem of building a practical, positive remote-control system has plagued amateurs for years. Some of the suggested schemes used have been very complex while others have been simple, but not foolproof. The technique applied in the design described here uses the telephone line. Positive and automatic control in the event of line failure are the basic assets of this unique system. Don't be alarmed at the apparent complexity of the schematic, **fig. 1**, as all the IC gates are contained in five, inexpensive SN7400 IC packages. Total cost of the circuit for ICs is less than $5.00 and a well stocked junk box will supply the balance. Although the circuit can be built on peg board, a custom made printed-circuit board is available and is recommended for simplicity.* See **fig. 3** for a suggested parts layout.

construction

A small chassis was made from sheet aluminum and bent in a vise to form the chassis shown in **fig. 2**. The power supply and relays are mounted on the chassis for convenience. A small equipment box could have been used but one was not available at the time I built the system. It is suggested when you choose your chassis that you keep the top and bottom open to aid in troubleshooting. A coat of white paint on the front panel and instant press-on lettering may be used to dress it up.

The model illustrated is set up for a

*Printed-circuit boards and nonlatching relays are available from Circuit Board Specialists, 3011 Norwich Avenue, Pueblo, Colorado 81008. Circuit boards are $5.50 each; nonlatching relays, $2.00 each.

latching function—it turns our repeater off. If you happen to have a stepping relay, this is fine, but I didn't, so another SN7473 was mounted in a socket off to the side and wired in as shown in the alternate circuit, **fig. 4**.

operation

An operating function, such as shutdown or turn-on of a remote station, requires a specific number of rings on the telephone, hang up, wait a specific length of time, then another specific number of rings and the function will be carried out.

In this discussion I will use 3 rings, hang up, wait 20 seconds, another 3 rings and hang up. Any combination of number of rings may be used so long as the total is less than nine. By examining the decoder (U2) you can see that it can be very easily programmed by the moving of only two jumper wires to the various outputs of U2.

Depending upon the type of relay used at relay K2, a momentary, latching or stepping function can be obtained. Relay K1 is used for validating the phone line. The remote station keying voltage can be taken through the contacts of this relay. If the phone line is interrupted, the transmitter cannot be activated.

circuit description

The phone line must be connected with the polarity as shown in the schematic. To assure the phone line being operational, the negative voltage is passed by CR1 and R1, charging C1, which will store the voltage during brief interruptions such as when the phone is ringing. The negative voltage applied to the gate of Q1 causes it to cease conducting. A high plus voltage will appear on the drain of Q1. This voltage is passed to the Q3 base by R4, causing Q3 to conduct and close relay K1. For repeater control K1 is connected in the push-to-talk or transmitter keying line so that it must be closed in order for the repeater to operate.

Although an explanation of the ring-counting circuit is a bit complicated, operation is actually very simple. In the ring circuit you are only interested in the ac signal. Capacitor C2 is used to block the dc and pass the ac signal to the rectifiers. The negative dc voltage is smoothed by C3 and C4 and applied to the gate of Q1, causing a rise in voltage on the drain of Q1. As soon as the ring signal disappears, R7 pulls the gate of Q1 to ground, resulting in full conduction of Q1, causing pin 14 of U1 low. Each shift from high to low on pin 14 will cause U1 to toggle once.

Therefore, the first ring will step the decade counter (U1) one count. The decoder-driver (U2) has now moved off zero position to its first count. When U2 leaves zero, pin 16 goes to +5 volts. This high is fed to pins 1 and 2 of gate 17 (U8A), and pin 3 goes low, causing the ready light to go out. Also, pins 4 and 5 of gate 18 are low, causing pin 6 to go high, turning on the start light indicating that a function has started. The high from gate 18 pin 6, is also fed to U7, pin 14, providing a set signal for flip-flop 1 (U7A), as well as providing the clear voltage to both flip-flops at pins 2 and 6.

To back up a little, to U2 pin 16, +5 volts is also fed to the emitter of Q4. This is the timing circuit of the telephone controller which provides the 20-second timing necessary for all phases of the function. The time is controlled by the RC network R11 and C5.

This all happens on the first ring of the telephone. The second ring simply advances the counter to count two. The third ring, the one that does the business, advances the decoder driver, U2, to count 3 position. The resultant low on pin 9 (count 3) of U2 is seen at pins 1 and 2 of gate 3. Gate 3, pin 3, will provide a high enabling voltage to pin 12, gate 1. Now, assuming that the telephone quit ringing on the third time, pin 12, gate 1 will remain high. After 20 seconds the unijunction transistor, Q4, will have charged up and fired a voltage spike out B1. This spike is fed to pin 13, gate 1. The output of this gate goes low, clocking FF1 (U7A) at pin 1. The Q output of FF1 goes high,

applying a set voltage to FF2. At the same time the phase 1 lamp comes on indicating that the first requirement of a function has been accomplished. Pin 11, gate 1, is also fed to gate 16, pin 2, which is used to reset the whole system if the telephone fails to ring again within the next twenty seconds.

When phase 1 has been successfully completed, the telephone must ring three more times to complete the function. The fourth and fifth rings will simply step the ring counter two more counts. On the sixth ring the phone must be hung up, leaving U2, pin 11, (count 6) low. This low is seen on pins 4 and 5 of gate 4, providing a high to pin 9 of gate 2. Twenty seconds after the last clock pulse from Q4 another clock pulse is sent to pin 10, gate 2. Gate 2, pin 8, goes low, causing FF2 to toggle. Pin 9 of FF2 goes high, causing the phase 2 lamp to turn on. This action also turns on the relay driver, Q5, closing the function relay, K2. Transistor Q5 will remain turned on until the third pulse is received from Q4. This time the clock pulse is fed to pin 2, gate 14. Pin 1 is now high since it is controlled by the $\bar{Q}$ output of FF2. The pulse is fed through gate 14 and 15 to reset U1. This restores U2 to zero, clearing FF1, FF2 and turning on the ready light. Thus, one complete function has been performed.

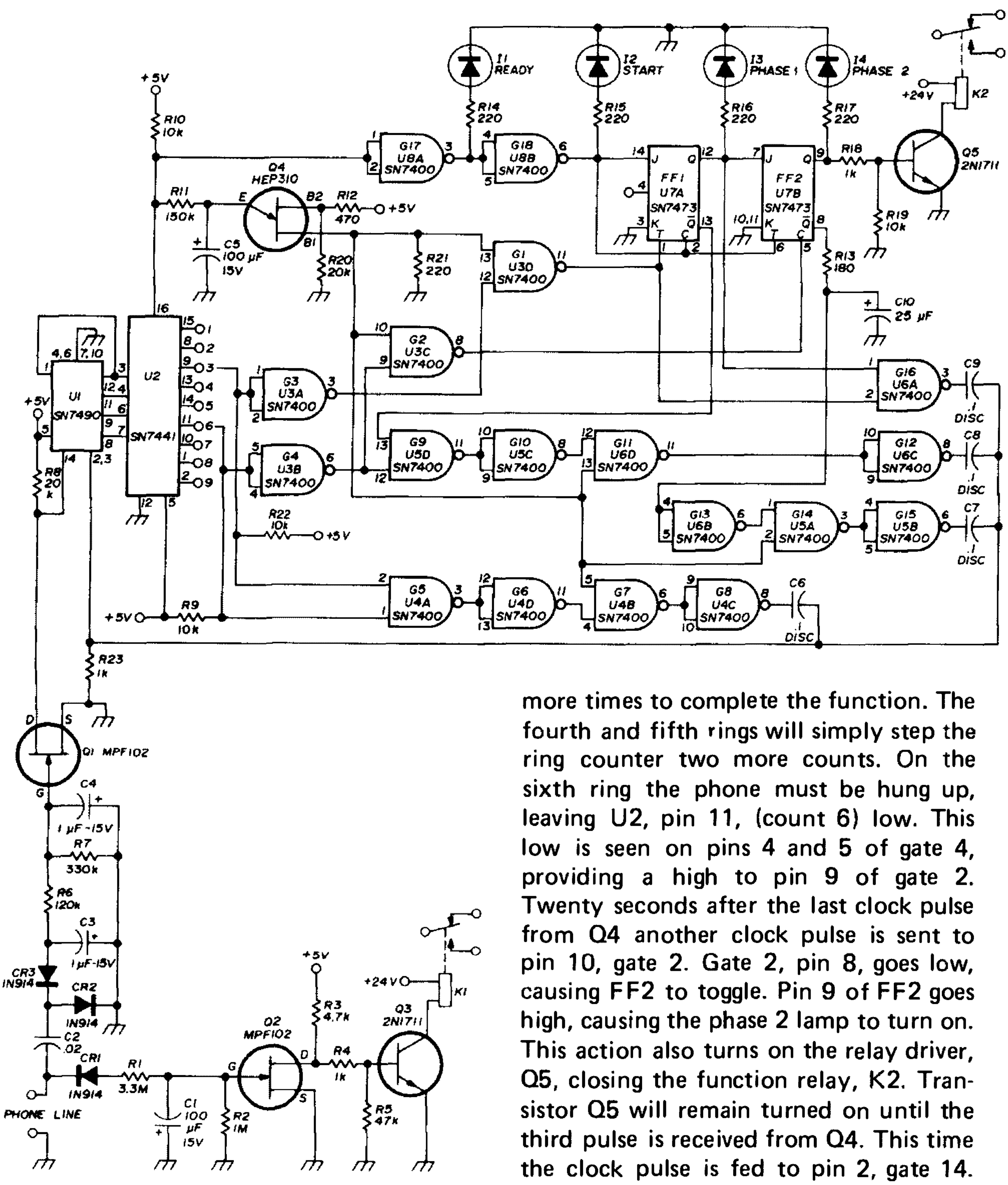


fig. 1. Schematic diagram of the automatic phone controller. All gates are SN7400 or equivalent. Relays K1 and K2 are sensitive dpdt relays with 8000-ohm coils. Resistor R11 is selected for the desired time setting. Resistors R14, R15, R16 and R17, in series with the LEDs, are adjusted for proper light output.

At this point you may be asking what the rest of the gates are for? Well, there must be a means of discriminating between a valid function and an incoming phone call to your wife or teen-age luck or accident they call back twenty seconds later and let it ring any number of times other than three or six. Then the entire system will be reset through gates 13, 14 and 15.

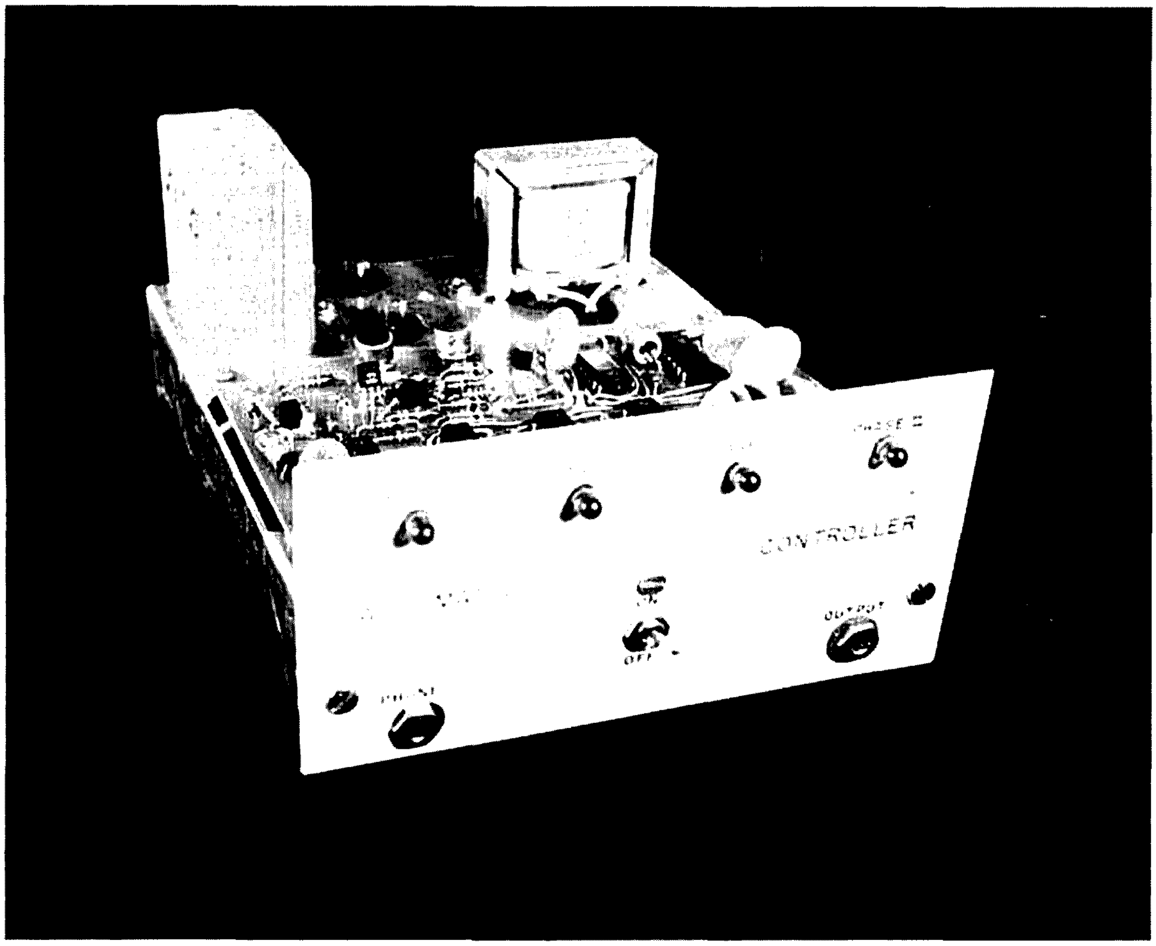


fig. 2. Overall view of the completed controller. Small jacks on the front panel are used for connections to phone and output. (Photo by Russell McGee, WBØGSU.)

daughter. A few instances of an invalid function would be:

1. Ring three times and hang up. The phone, not ringing again for the next 20 seconds, would allow the system to reset through gates 1 and 16.

2. The phone ringing any number of times except 3 or 6 then hung up, the system will be reset through gates 5, 6, 7 and 8.

3. The last possibility is a remote one but nevertheless it must be considered. In this case someone lets the phone ring three times and hangs up. By sheer stroke of

This concludes the operation of the controller. As far as I know I have thought of all possibilities of mis-control and have provided protection against it. If, by accident, someone does get your function-control code it is a very simple matter to change the sequence to another code, such as ring 7, hang up and ring 2 more times. As you can see, there are 36 different combinations of sequences. If that isn't enough you can change the value of R11 and C5 for a different time duration. Needless to say, it isn't necessary to describe the multitude of combinations you can come up with using this arrangement.

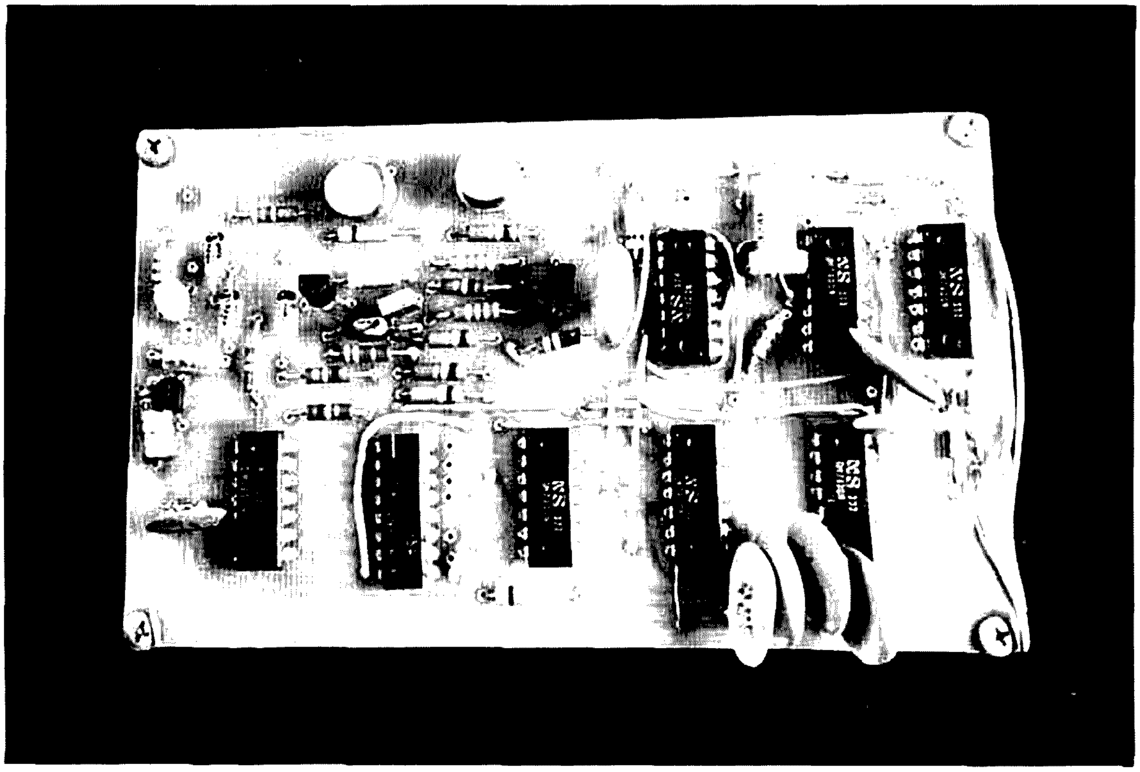


fig. 3. Component layout. The two transistors at the top are Q3 and Q4. The input and output connections are on the bottom; wires run to the light-emitting diodes located on the front panel. (Photo by Russell McGee, WBØGSU.)

troubleshooting

A high-impedance scope or a vacuum-tube voltmeter will be required in the event of trouble. However, the light-emitting diodes connected to each stage will generally suffice for troubleshooting.

After construction is completed and voltage is applied to the circuit check for the following: the ready light, I1, should be on and all others off. Pin 16, U2, should be low, pins 9 and 12, U7, should be low and pins 8 and 13, U7, should be high. Connect the positive terminal of a small 9-volt battery to ground and the negative side to the input. Relay K1 should close and remain closed for about 30 seconds after removing the 9-volt battery.

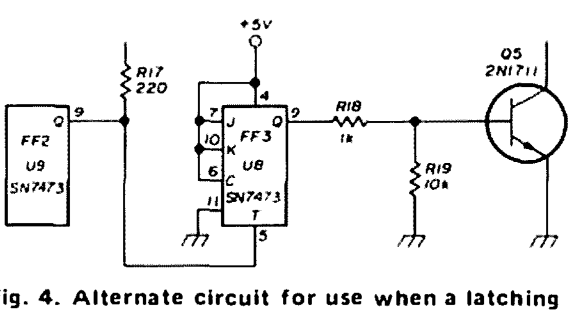


fig. 4. Alternate circuit for use when a latching function is required.

Next, apply the 9-volt battery lead momentarily to the junction of CR3 and R6. Approximately 2 seconds after removing the battery lead I2 should light, indicating a function has started. Make and break this connection, allowing about 2 seconds between pulses for the correct number that corresponds to the first number of rings that you have programmed for. About 15 seconds later I3 should light, indicating that the first phase of the system has been satisfied. Now make and break the battery connection in the same manner for the second combination of rings; in approximately 5 seconds I4 will light, indicating that the sequence has been completed and the function has been carried out.

In the event of difficulty check base 1 of the unijunction transistor with a scope for the pulse that controls the system. If the pulse is being generated, then trace the pulse through the system to determine which gate or flip-flop is not responding.

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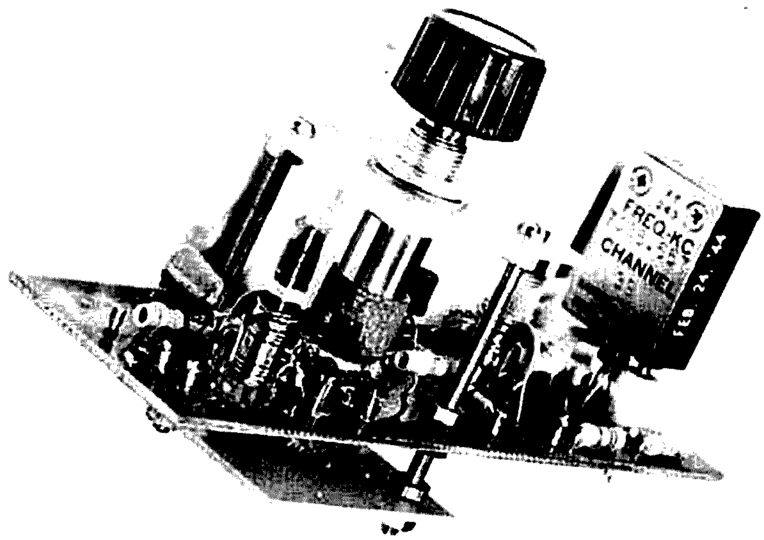


tuned very low-frequency converter

Novel tunable inductance provides basis for tuned VLF converter that covers wide frequency range without bandswitching

Guenter Ruehr, OH2KT, Porintie 1 C 35, SF - 00350 Helsinki 35, Finland

There is considerable interest in the very-low frequencies, particularly for the WWV transmissions on 20 and 60 kHz, and for operation on the no-license band around 1750 meters.[1] One of the big problems in building receivers or converters for this part of the spectrum is in constructing a variable tuned circuit which will cover a substantial portion of the desired frequency range. Assuming that the desired band extends from 10 to 150 kHz, with a ratio of the corner frequencies of 150:10 = 15, the tunable component must have a variation of $15^2 = 225$. Since this cannot be accomplished with conventional variable capacitors or inductors, the frequency range has to be divided into a number of sub-bands or the tuned circuit is eliminated altogether. The latter is done in most VLF converters – they are untuned.

tuned circuit

There is, however, a novel method of inductive tuning which will cover the required range.[2] This method makes use of a toroidal ferrite core which is magnetically biased by a pair of small permanent magnets as shown in **fig. 2**. By rotating one of the magnets with respect to the other, the amount of flux penetrating the toroid is varied, changing the ferrite's permeability and thus, the inductance. It is interesting to note that maximum flux penetration and minimum inductance occur when like poles are opposite one another.

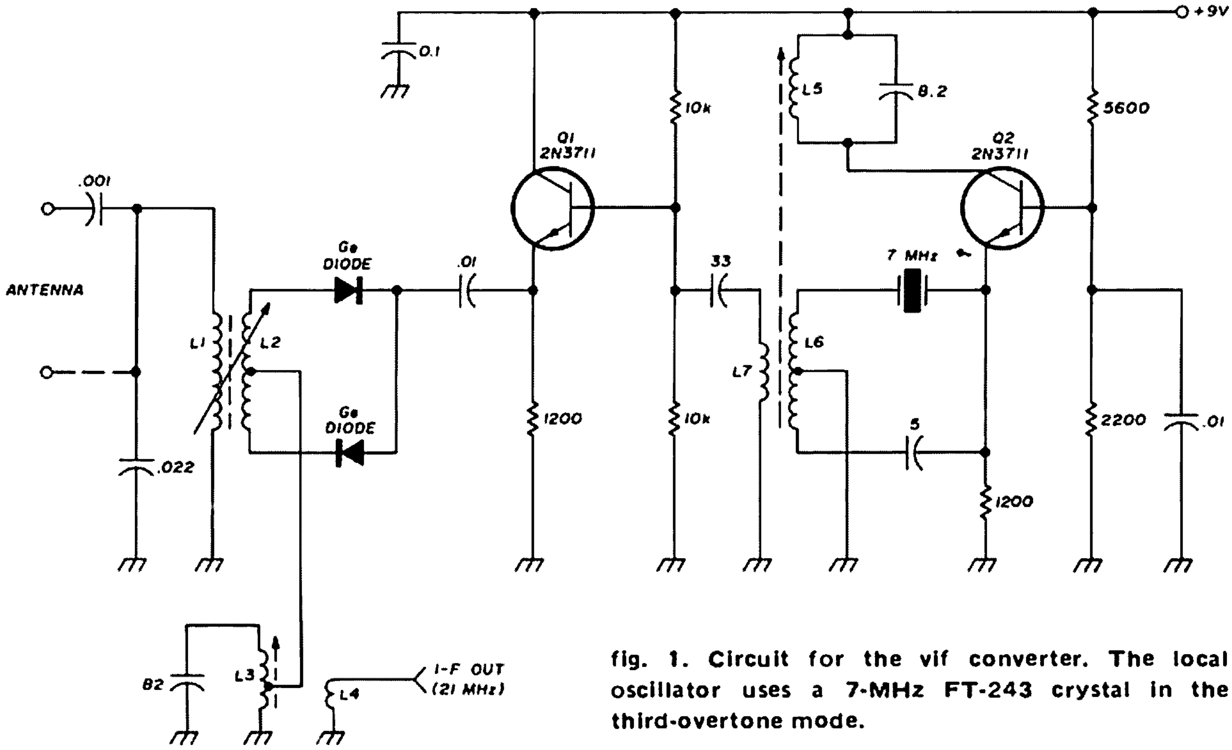
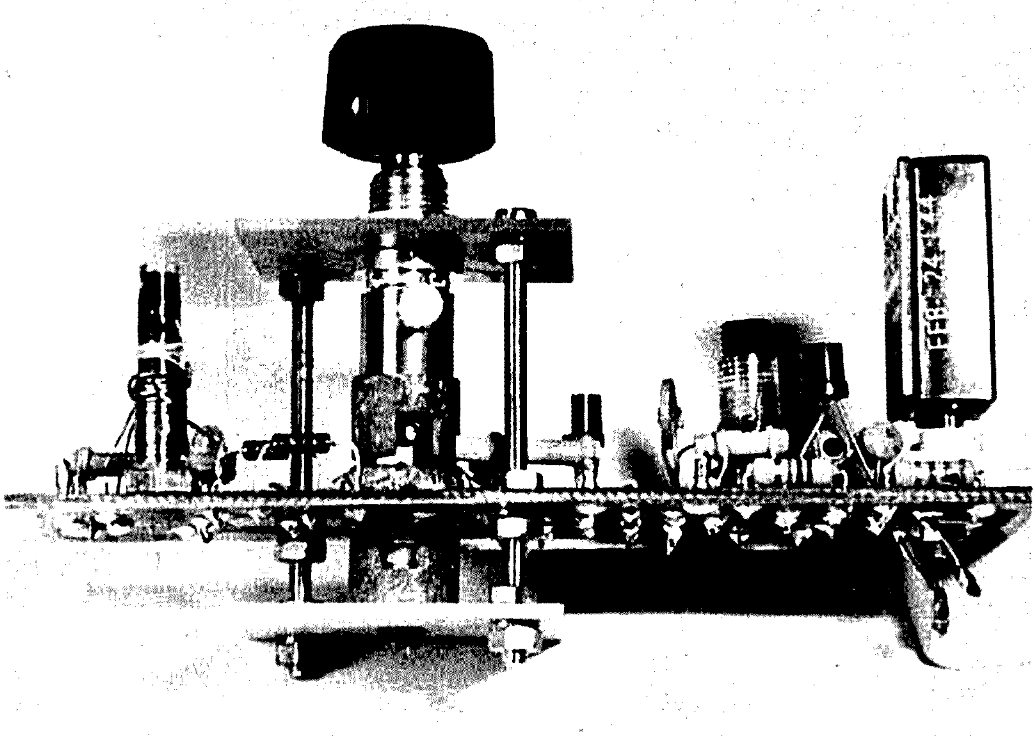


fig. 1. Circuit for the vlf converter. The local oscillator uses a 7-MHz FT-243 crystal in the third-overtone mode.

L1,L2	magnetically tuned inductor (see text)	L5	15 turns no. 20 on ¼" (6-mm) slug-tuned form
L3	10 turns no. 20 on ¼" (6-mm) slug-tuned form, tapped 5 turns from cold end	L6	4 turns no. 20, center tapped, around cold end of L5
L4	2 turns no. 20 around cold end of L3	L7	2 turns no. 20 around cold end of L5

The two magnets used to bias the toroid inductor are of the *button* type with a half-inch (13-mm) outside diameter. The outside diameter of the toroid is also ½-inch. The whole tuning assembly is built around the bushing and shaft of a discarded potentiometer. The particular toroid core I used required 100 turns of



Closeup of the tunable vlf converter showing magnetically-biased tuning inductor.

stranded wire for an inductance variation of 100 μH to 12 mH (a 120:1 range). However, ferrite cores with higher permeability would require fewer turns. Measured Q values for my inductor were around 50 for frequencies between 10 and 150 kHz.

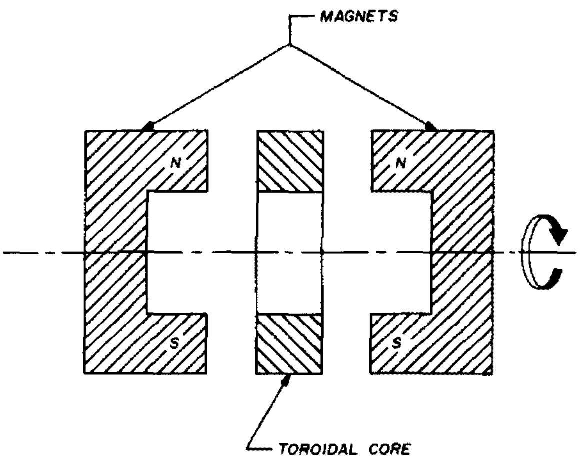


fig. 2. Method for magnetically biasing a toroidal ferrite core with two small button magnets. This technique provides an extremely wide inductance tuning range.

converter circuit

The circuit for my VLF converter, which has an output on 15 meters for use with a communications receiver, is fairly conventional as shown in **fig. 1**. The antenna is coupled directly to the hot end of the tuned circuit (or through a capacitor to provide a degree of matching to long antennas). The mixer uses a matched pair of germanium diodes and the local oscillator uses a FT-243-variety crystal in the third-overtone mode. To obtain third-overtone oscillation at 21.000 MHz, choose a crystal with a fundamental frequency a few kHz above 7 MHz. By simply changing the crystal frequency and the oscillator and i-f output coils, output can be changed to any desired frequency band.

references

1. J.V. Hagan, WA4GHK, "A Crystal-Controlled Converter and Simple Transmitter for 1750-Meter Operation," *QST*, January, 1974, page 19.
2. N.H. Brown, "Miniature Wide-Range VLF Tuner," *Electronics World*, July, 1971.

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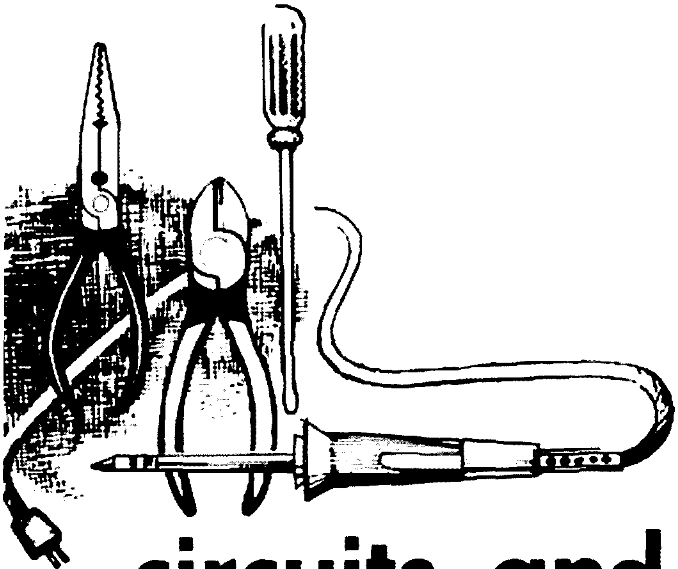


circuits and techniques

ed noll, W3FQJ

W3FQJ goes solar power

On April 15, 1974, the solar-powered QRP station at W3FQJ made an initial radio contact with Hank Brazeal, WB4ZXJ, in Birmingham, Alabama on 15 meters. The transceiver was the five-watt PEP Ten-Tec Argonaut. The same evening W3FQJ checked into the RF Hill Amateur Radio Club's 10-meter net.

The essential units of the solar-powered station are the Argonaut transceiver, charging panel and 5.5-ampere-hour motorcycle battery (fig. 1) and a roof-mounted light-energy converter, (fig. 2). The 5x2½x5½-inch (12.7x6.6x14-cm) battery is positioned behind the charging panel for normal operation. Such batteries can be purchased at local motorcycle shops or by mail from one of the auto accessory houses or from Sears. The Sears batteries are shipped with a dry electrolyte which must be added to the battery along with water according to the instructions. In most shops the electrolyte is added when the battery is purchased.

The Spectrolab light-energy converter has a rating of 12-volts at 0.3 amperes. When the solar panel is so operating it is supplying 3.6 watts. This is an average figure. At dawn and dusk and during dark, overcast days the power level is significantly lower. When the rays of the sun are striking the panel directly, the delivered power is slightly more. Nevertheless, it will make available more than enough electrical energy for a very busy five-watt QRP station.

The solar panel can be operated as a continuous float-voltage charger or it can be operated whenever you wish to recharge the battery. As I am writing this column in the early morning hours of a high-overcast day the float-voltage connection is supplying 50 milliamperes.

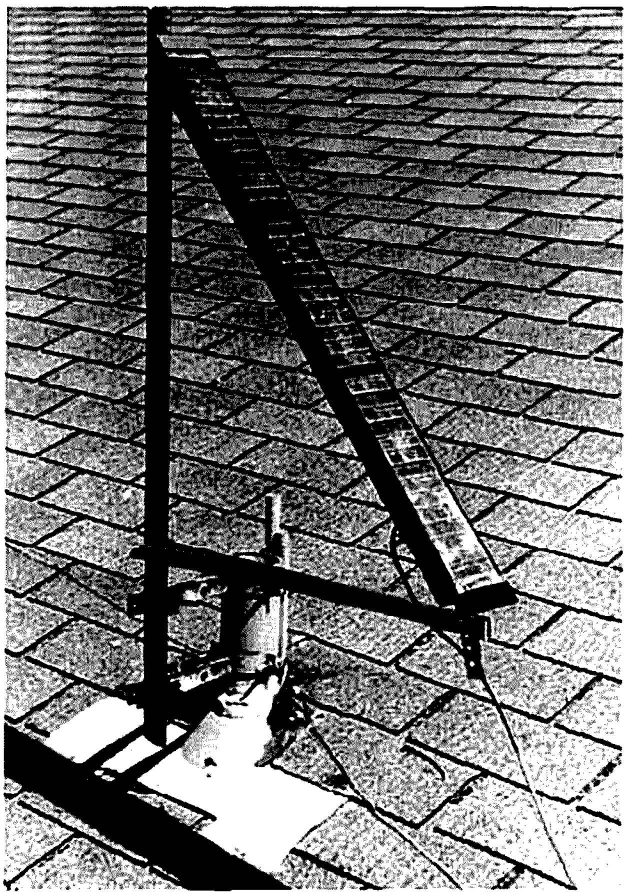


fig. 1. Solar power converter used by W3FQJ is mounted on the roof. Details of the mount are discussed in the text.

Yesterday in bright sunlight a series resistor was inserted to keep the trickle charge current below the 100 mA level. As set up now the station plan is very conservative and would provide continuous operation for normal back-and-forth amateur chatter. Over a period of time I will learn the limits of the system, pressing high-powered transmitters into operation. Such is not conducive to the furthering of corporate empires.

charging panel

Several components are needed for the charging panel. Refer to the schematic diagram and parts list shown in fig. 3. About 50-feet (15-meters) of two-conductor cable (number-12 or -14 conductors) connects the solar polar converter to the charging panel. The panel itself supports a dc milliammeter, a 1-ampere diode, CR1, and a number of binding posts that permit ease in monitoring and experimentation.

fig. 2. Basic solar-powered QRP station used by W3FQJ uses Ten-Tec Argonaut transceiver, solar power converter (fig. 1), solar control panel (fig. 4) and 12-volt motorcycle battery.

Reports will appear in this column from time to time.

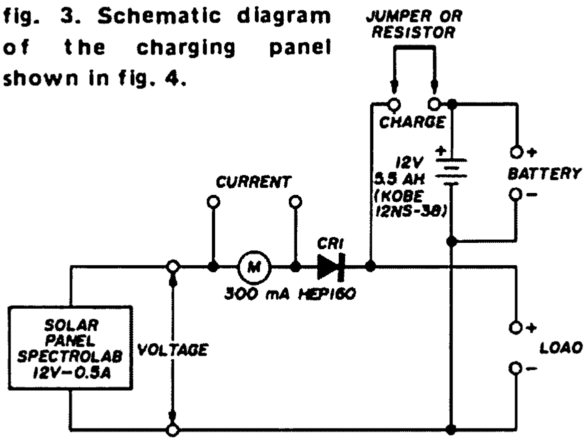
solar power cost

Today solar electric power is expensive. The small 3x39-inch (7.6x99.1-cm) unit I am using lists for more than 150 dollars. Much about it is handcrafted and as yet systems have not been adapted to mass production. Unfortunately, it must be the American citizen that is obliged to set up a widespread clamor for solar power. One cannot anticipate that the oil institute, the electric power industry or the atomic energy bureaucracy will do too much to further the cause of this non-polluting method of power generation. Such power would offer a degree of independence for many homes, small businesses and small industrial plants.

The two binding posts at the top right of the front panel, fig. 4, can be used for monitoring the voltage delivered by the solar panel. The meter to the left reads the actual current being delivered by the solar cells. The two binding posts underneath the meter permit the insertion of a jumper if you want to take the meter out of the circuit for those experiments where you want to draw maximum current under the condition of bright sunlight directly striking the cell surface.

The battery terminals are at the lower right. In a continuous service application

fig. 3. Schematic diagram of the charging panel shown in fig. 4.



the 12-volt device being powered is connected across these two terminals. When the battery is to be charged, a jumper is connected between the two terminals labeled *charge*. If you wish to limit the charge to a specified current level under bright conditions, a resistor can be connected between these two terminals. Often, under very bright conditions, I shunt a 100-ohm, 5-watt resistor between the terminals.

The two load terminals at the bottom center of the panel permit you to supply energy directly from the solar panel to a 12-volt device. In this application the battery is disconnected completely by making certain there is no jumper or resistor connected across the two charge terminals.

Diode CR1 is an important part of the charging system. The actual charging current declines as the battery reaches full charge. This is to be anticipated because the charging battery attains a voltage ever closer to the voltage of the charging source. However, it is possible, especially when the battery climbs to full charge voltage, that the impinging light will not be great enough to maintain the charging source voltage above the level of the battery voltage. Without the diode in the circuit, the battery would then discharge into the solar source. This is avoided because under the condition of high battery voltage and low charging source voltage the diode is reverse biased (cath-

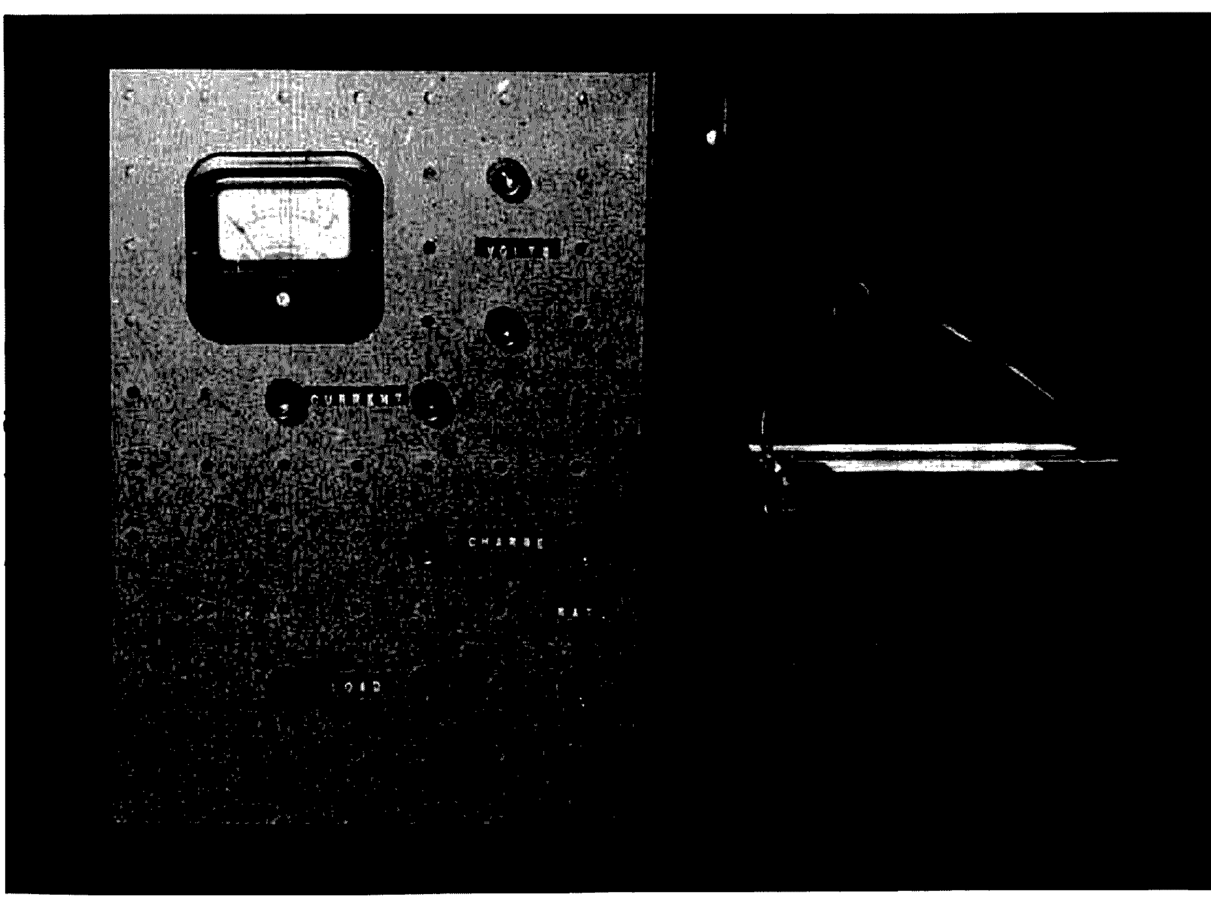


fig. 4. Solar control panel and 12-volt battery used by W3FQJ. Control panel is mounted on 8x11-inch (20.3x27.9-cm) piece of Masonite peg-board. Circuit is shown in fig. 3.

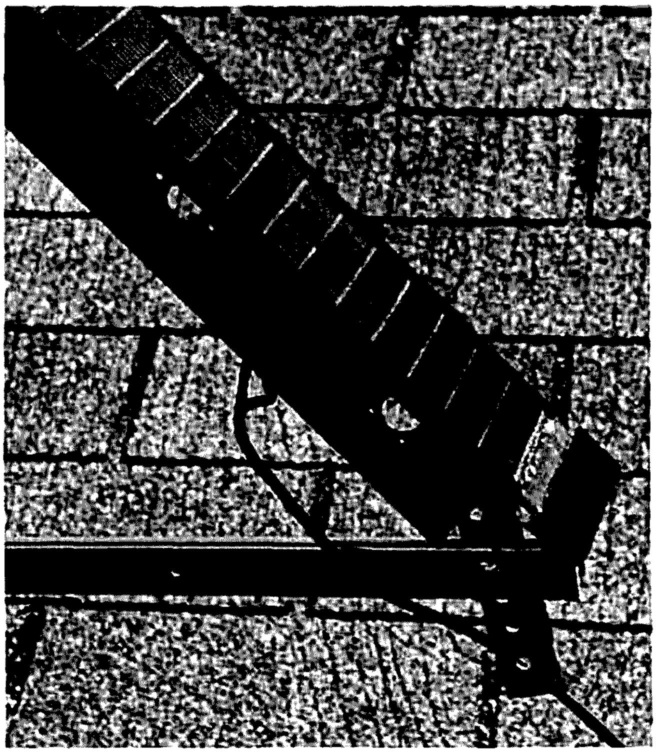


fig. 5. Closeup, showing tilt angle adjusting mechanism for the solar energy panel.

ode connected to positive side of battery and anode connected to the positive side of the charging voltage).

roof mount

The basic physical support for the solar panel is the vent pipe on the roof of my house, fig. 1. At one time the two standoff brackets supported an antenna mast. A shortened 5-foot mast section and a homemade bracket arrangement supports the solar panel. It was arranged to permit ease in experimenting with the tilt of the panel. A U-bolt permits the top of the panel to be moved up and down the mast. At the bottom of the panel a flat piece of aluminum with a series of holes permits easy accommodation of various tilt angles.

The recommended tilt for the panel corresponds to the latitude of your station (degrees north [or south] from the equator). My station is reasonably near the 40° north latitude line. This figure refers to the angle of tilt away from the horizontal as shown in fig. 6. The nearer you approach the equator, the nearer the optimum mounting angle approaches a horizontal position.

In setting up the mounting arrangement the solar panel becomes the hypotenuse of a right triangle, fig. 7. Sine and cosine functions can then be used to determine the vertical and horizontal sides. The overall panel length is 39 inches (99.1 cm); this becomes the length of the hypotenuse. Therefore, the horizontal and vertical side dimensions become:

$b = h \sin a$	$a = h \cos A$
$b = 39$ inches (99.1 cm) $\times \sin 50°$	$a = 39$ inches (99.1 cm) $\times \cos 50°$
$b = 30$ inches (76.2 cm)	$a = 25$ inches (63.5 cm)

The vertical side is an appropriate section of the 1¼-inch (32-mm) mast section. The horizontal support is two 1¼-inch (32-mm) aluminum strips bolted together but separated where they wrap around the mast section and where they connect to the short length of aluminum strip that permits the adjustment of tilt angle.

battery data

A 5.5-ampere-hour capacity battery provides a conservative and well-regulated source for the 5-watt Argonaut transceiver. Transceiver specifications suggest a 12-volt, 1-ampere source although peak current demand by the transceiver is less than this value. At any rate, such a battery would supply the unit for continuous overnight operation. In fact, the battery could supply almost a continuous demand of 1 ampere for 4 to 5 hours.

Based on a 20-hour, 5.5-ampere-hour rating, the continuous current demand for this period of time would be:

$$I = \frac{5.5}{20} = 275 \text{ milliamperes}$$

A charging current of the same value would recharge the battery in the same amount of time plus additional time, depending upon the charging efficiency of the battery.

In normal amateur applications deep discharge of the battery and continuous high-current charging are not necessary. Once the battery is fully charged a float

charge arrangement is quite adequate. If you assume a charge rate of 1/4 to 1/6 of the rated current value, you are considering a minimum charge current in the 45- to 70-mA range. In the set-up described this level of current and higher is readily available for hazy-bright days. A generous, up to 300-mA, current is available during bright sunny days. Even on a high overcast day the solar panel supplies current at the low end of the range. Charging current is limited for dark days but the battery capacity is adequate, even for the very active QRP operator.

solar power QSO record

Correspondence from Edgar Janes, G2FWA, disclosed two solar-powered CW contacts made on October 27, 1954. Initial contact was made between G3HMO (solar-powered station) Buckingham, England and G5RZ, Leighton Buzzard, England. Here is an account from the December, 1954, issue of *Short Wave Magazine*:

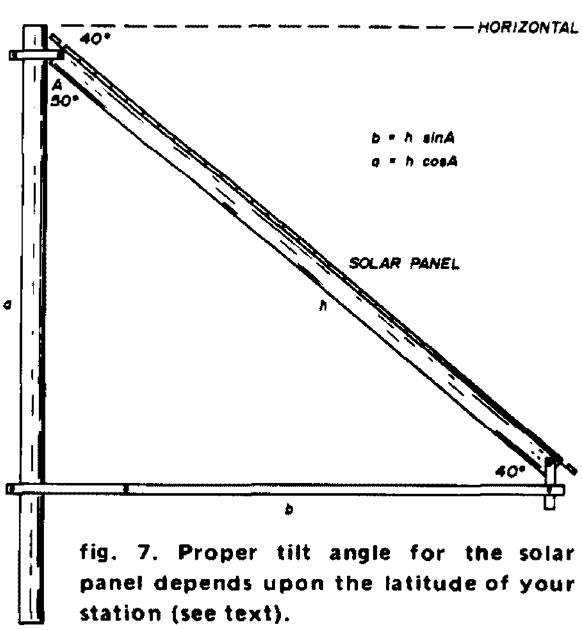


fig. 7. Proper tilt angle for the solar panel depends upon the latitude of your station (see text).

"Fortunately, the sun was shining fairly strongly, and the photo-electric cell battery was giving ample output – about 2 mA at 4 volts – to energize the transmitter . . . Two-way CW contacts (were made) with G5RZ (Leighton Buzzard, 15 miles) at 1505 gmt and with G3IYX (Bradwell, Bucks, 7½ miles) at 1515 gmt. The daylight powered transmitter signals on 1820 kHz were reported as RST 559 in Leighton Buzzard and RST 569 in Bradwell.

These contacts, though pre-arranged, were initiated on the transistor trans-

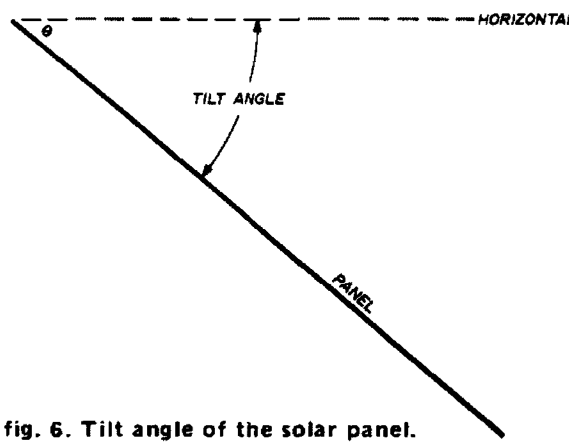


fig. 6. Tilt angle of the solar panel.

mitter running on the photo-cell battery alone and were carried through under normal band conditions, with quite troublesome interference on the frequency."

Thank you, Edgar, and Austin Forsyth, G6FO, Managing Editor of *Short Wave Magazine*.

My contact with WB4ZXJ on April 15, 1974, is degraded to the first solar-powered ssb contact, not arranged. Maybe? The paramount question, however, is why we have waited so long (except for satellite communications) to take advantage of this limitless source of energy.

new applications

More small companies are now producing small solar energy converters. Zurn Industries* sells two models with peak ratings of 1.5 and 6 watts. These two 12-volt models are covered with a transparent waterproof and detergent-proof coating. A major application of the Zurn models is for marine batteries. Under the condition of a week of normal daylight, sufficient charging power is available that

*Zurn Industries Incorporated, 5533 Perry Highway, Erie, Pennsylvania 16509.

a bilge pump can pump 600 gallons of water with the small energizer and 3000 gallons with the larger model. Such a device is particularly useful for maintaining battery charge during long idle periods, as during winter storage of a small boat.

Another *active organization*, Solar Energy Company,† sells solar power converters as well as wind-powered electrical sources. Their devices find application in vhf/uhf repeaters, microwave relays, wire and wireless telephone systems, TV translators, monitors, offshore platforms, data buoys, railroad signals and controls, plus traffic and security systems. They also make a unit that maintains the charge on batteries used to power farm electric fences.

Why haven't such devices been used to supply a portion of the power needed in the average household? Needed is a solar heater for the home with its electrical blowers, fans and circulators powered by a solar energy converter. At least there should be a home furnace capable of using a variety of fuels with its electrical accessories powered completely by battery and solar power converters.

experimental approaches

There are several avenues of experimentation. What are the operating limits using direct drive to 12-*volt* devices and no battery? Some voltage-regulator device would be required. How bright would it have to be to provide operation of a 12-volt zener diode? How much longer a period of operation would be feasible using the solar source for 9-volt regulated operation? What are the limits of a particular solar panel in terms of higher power demand and keeping a higher capacity battery fully charged relative to your normal operating schedule?

Connecting solar panels in parallel increases current capability. How many panels would be required to provide adequate charging current for a high capacity battery system? How many panels and what battery capacity would provide enough power to match normal operating time for your 200-watt sideband transceiver? What advantages are to be obtained from series-parallel groupings of solar panels?

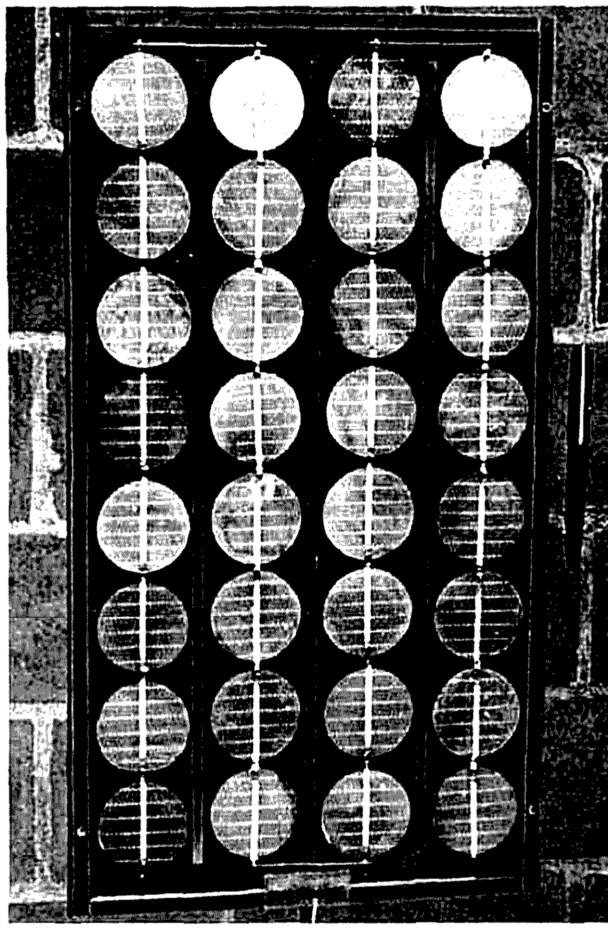
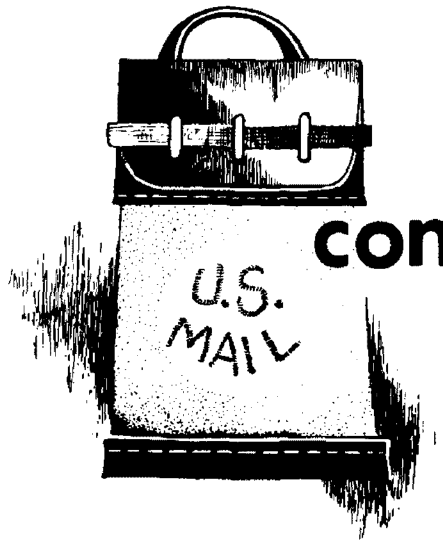


fig. 8. 12-volt, 500-mA solar panel (courtesy Solar Energy Company).

What additional capacity can be gained by using a mechanical mount that would permit you to chase the sun across the sky? Would an equatorial mount and clock-drive be practical? How would you hold the weight of the assembly down? What percentage of the derived power would be needed to power the drive system? Perhaps drive power could be conserved by changing positions only once each hour, or half-hour. I will be discussing these and similar subjects in the months ahead.

ham radio

†Solar Energy Company, 810 18th Street, NW, Washington, D.C. 20006.



comments

coherent fsk RTTY

Dear HR:

I would first like to commend Steve Maas, K3WJQ, for his article bringing modern communication theory into the area of amateur RTTY in the June, 1974, issue of *ham radio*. However, I feel that a few areas need comment. First, the numbers in table 1 (page 32) for coherent FSK probability of error are slightly in error. The published data is based on a shift between the mark and space frequencies of $f_s \approx 0.7/T$, where T is a baud time.[1] For 60 wpm RTTY, T = 0.022 second, resulting in an f_s of only 32 Hz. It can be shown that this shift results in the lowest probability of error for coherent demodulation of FSK. However, for many practical reasons, this is too narrow a shift, so for standard 170- or 850-Hz shifts, that column should be amended to

input snr	probability of error coherent FSK
0 dB	0.16
2 dB	0.10
3 dB	0.08
6 dB	0.02
9 dB	0.002

The probability of error, as well as the Rayleigh and Rician probability density functions, are derived assuming no fading, with white noise having a Gaussian density function at the inputs of the mark and space filters. This is sometimes a good assumption, but primarily at vhf and above. Since most RTTY activity is in the high-frequency spectrum, this white Gaussian noise assumption needs examination.

The National Bureau of Standards has attempted to model atmospheric noise.[2] They have found that the performance of an FSK demodulator can be several orders of magnitude worse than calculated by the white Gaussian noise assumptions, and this study didn't include QRM, which seems to be the most prevalent type of "noise" in our crowded bands.

The high-frequency spectrum is time varying with frequency-selective fading (QSB). This fading results in a tremendous degradation in the probability of error performance. Fading is thought to be due to constructive and destructive combining of various received signals that have been reflected from different layers of the ionosphere, resulting in a received signal with varying amplitude and with a random phase angle. So, while the transmitted phase may be constant, because of the varying time delays through the propagation path, the received signal can have any phase. It appears that demodulation schemes which require phase information, such as coherent FSK, are degraded *more* than noncoherent schemes.[3]

By requiring that a PLL (fig. 4 in K3WJQ's article) remain close to the previously transmitted phase while the

1. Schilling and Taub, *Principles of Communications Systems,* McGraw-Hill, New York, 1970.
2. Alyce M. Conda, "The Effect of Atmospheric Noise on the Probability of Error for an NCFSK System," IEEE Trans. on Communications Techniques, September, 1965, page 280.
3. G.L. Turin, "Error Probabilities for Binary Symmetric Ideal Reception Through Nonselective Slow Fading and Noise," Proceedings of IRE, September, 1958, page 1603.

other channel is being transmitted would result in a very long time constant in the lowpass filter. This would also have the effect of increasing the time it takes for the PLL to lock up. PLLs in amateur demodulators have usually had the additional problem of locking on to a strong nearby interfering signal. Finally, there is evidence that the use of two oscillators at the transmitter end, which results in discontinuities in the phase of the transmitted signal, will have poorer performance than shifting the frequency of a single oscillator.[4]

Most mathematical analysis of systems like the one described in K3WJQ's article ignore the effect of one bit on the next. This problem is called intersymbol interference, and is found in any receiver having narrow bandpass filters. The degrading effects of intersymbol interference are extremely difficult to analyze, but they are of concern, as evidenced by the multitude of articles on attenuation and delay equalizers in the technical journals.[5]

The major causes of errors in an amateur's RTTY system thus seem to be those caused by fading, high atmospheric noise levels, interfering stations and intersymbol interference introduced by the propagation medium and receiver. Table **1**, while slightly in error in practice, does point up the fact that a small increase in antenna gain can tremendously improve the performance. Also, for those fortunate hams with forty acres and unlimited resources, space diversity antenna systems reduce the effects of fading. Other known techniques for removing the effects of interference or fading, such as frequency diversity, error correcting codes and spectrum spreading methods, aren't looked upon favorably by the FCC.

4. P.A. Belle and B.D. Nelin, "The Effect of Frequency-Selective Fading on the Binary Error Probability of Incoherent and Differentially Coherent Matched Filter Receivers," IEEE Trans. on Communications Systems, June, 1963, page 170.
5. A. Gersho, "Adaptive Equalization of Highly Dispersive Channels for Data Transmission," Bell System Technical Journal, January, 1969, page 55.

I hope that the above comments will stimulate thinking on the design of FSK demodulators. The articles, such as the ones referenced, should provide indications as to the directions a design should take. I hope that I have shown that while I am enthusiastically for more theoretical articles in the amateur magazines, they should be tempered with the idea that mathematical analysis of various systems can be, at best, extremely difficult. The only way that a majority of amateurs, including myself, will be convinced of a new method's effectiveness is by seeing a working model perform as well or better than existing models.

John Fehlauer, WA2WTL
Gibbstown, New Jersey

bequests amateur gear

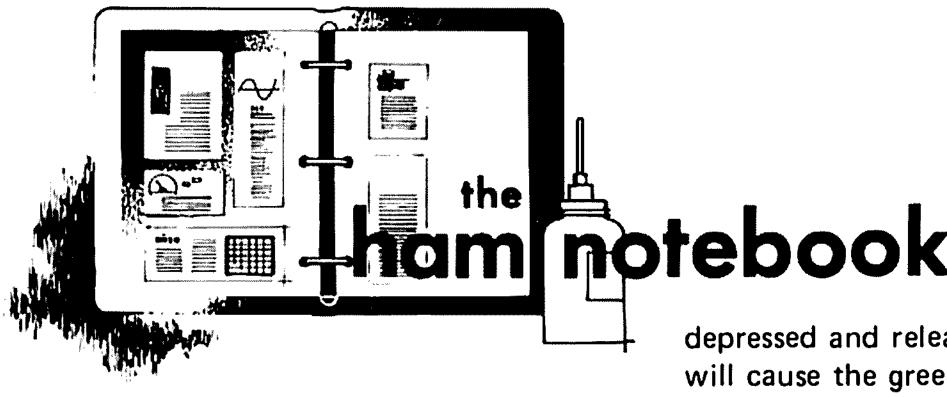
Dear HR:

Your March editorial was interesting and to the point regarding the disposition of a deceased ham's equipment. Specifically, it takes into account the fact that there is a big difference between *good* junk and *junk*. Too many of today's hams consider homebuilt equipment to be in the latter category, but I think I have the solution.

My will specifies that all of my equipment which can be categorized as amateur radio equipment and support parts shall be made available to the County Board of Education, with specific instructions that such items be sent to the local high school (which happens to have an amateur station). Further, I have specified that any taxes, fees and/or delivery costs be paid from my estate because I wanted to prevent any failure of delivery because of some cost or expense that the Board of Education might elect not to pay.

There are no strings attached, the high school will receive some valuable equipment and—hopefully—some youngster will learn something from the bequest.

Dean Young, W3FZ
Adelphi, Maryland



identification timer

Lengthy QSOs and time-consuming phone patches can lead to problems with the FCC if you do not identify your station every ten minutes. After becoming an Official Observer I became aware of times between ID—especially those of ragchewing stations and stations running phone patches. Operators in these two categories tend to let time ramble by without the proper 10-minute identification. To alleviate this problem at my station I set out to build a timer that would accurately reproduce a 10-minute time out, have a minimum number of parts, and could be assembled in one evening. The circuit is shown in fig. **1**.

This timer was designed around the Signetics NE555 integrated circuit. Any 9- to 12-volt dc supply can be used to run the timer. Once the power supply is turned on, the red lamp (*ident*) will light. The *run-test* switch is placed in the *test* position and the *reset* button should be depressed and released. A ground at pin 2 will cause the green lamp to light and the red lamp to extinguish—indicating that the unit is *timing*. With the values shown the *test* position will allow a time cycle of 2 to 3 seconds after which the lamps will change from green to red (*ident*). Now

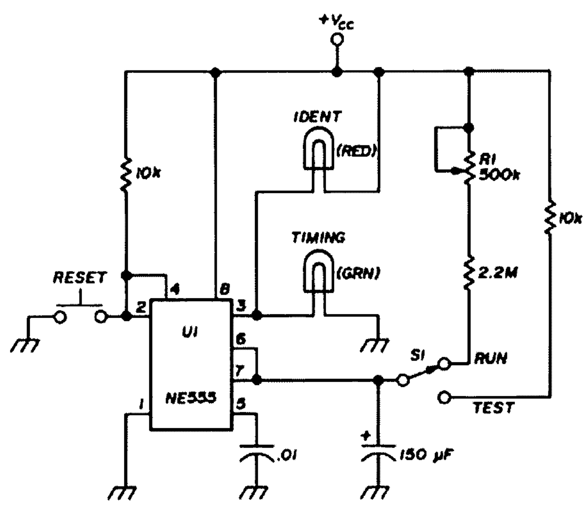
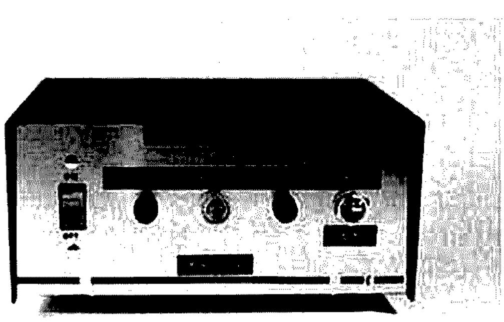


fig. 1. Simple ten-minute ID timer uses low-cost NE555 timer IC. Timing period, adjustable from 7 to 11 minutes, is set by R1.



place switch S1 in the *run* position. Again depress and release the *reset* button. The unit will time out between 7 and 11 minutes, depending on the setting of R1. I have R1 (500k pot) on my unit set to time out at 9 minutes. This allows a one-minute time frame in which to identify.

Even with ±10% variation from nominal supply voltage of 12-volt dc the timer retains a 9 minute, ±1 second, accuracy. If during a timing cycle you

wish to identify your station and return to zero time it is only necessary to depress and release the *reset* switch. It is recommended that the indicator lamps draw no more than 100 mA each so as not to strain the current-sinking ability of the NE555 (200 mA max).

The total cost to build this timer is under five dollars. Circuit layout is not critical and component minimization makes the unit very reliable. With a timer like this one next to your phone patch you don't have to worry about being cited for lack of a correct 10-minute ID.

Don Backys, K9UQN

spurious signals

The October, 1973, issue of *ham radio* (page 67) carried details of Yaesu FTdx560/570 vfo frequency ranges, and the incorporation of the 6.358.6-kHz suck-out crystal in recent production. This was summarized in the December, 1972, issue (page 69), based on JA1MP's letter earlier in the year. This modification eliminated a 28-MHz spurious, but does not appear to affect the clean lower-sideband phone signals appearing in the 14-MHz DX band in a manner similar to those reported from the Yaesu FTdx400.

W6PKK added the crystal to his FTdx560, then retrimmed traps L18 and L19 (but not L23). His 14270-kHz upper-sideband transmissions caused lower-sideband spurious on 14080 kHz. Trimming eliminated the spurious completely at my location. In the case of a W7 who installed the suck-out crystal ($5 plus COD postage), the 14197 lower-sideband spurious caused by the 14212 kHz upper-sideband transmission was not reduced; it presumably requires adjustment of the two traps as accomplished by W6PKK.

It appears that all Yaesu amateur equipment may have one or more traps which must be properly adjusted and the performance confirmed by listening tests.

A number of CW signals heard in the phone band may have been spurious from the CW band on 14 MHz, but were not all investigated when heard. One was on about 14345 kHz. Several RTTY signals around 14085/14090 kHz have also been heard in the vicinity of 14220 kHz, these being from HW100, SB101, SB400 and SB401 equipment, according to information received. Some of these sets appear to depend upon a "bandpass coupler" out of the first mixer to suppress any undesired second harmonics of the input frequencies, or of the resulting output frequency. If the bandpass coupler is not sufficient, suitable traps might be applied to the first mixer output or the second mixer input (if two mixers are incorporated) to suppress the undesired products.

Information should be accumulated on the spurious performance of all types of amateur equipment, and the necessary cure for any spurious emissions or responses, to eliminate unnecessary receiver interference.

Bill Conklin, K6KA

full-quieting meter for Clegg 27B

Many amateurs who use the Clegg 27B as a base rig have wanted a visual indication of incoming signal strength for peaking up mobile signals and for keying up repeaters. This can be simply accomplished as follows: Connect one end of a piece of miniature RG-74/U coaxial cable, center conductor to the orange wire on the squelch control and braid to the yellow wire of the squelch control. Run the RG-174/U coax back to the hole near the antenna connector and connect it to a miniature 1/8-inch female connector. For the meter, which is connected to the coax with a matching 1/8-inch male connector, I used a 0-500 microampere meter. A 0.01-μF capacitor across the meter will smooth out any needle movement. Make sure the cable braid is not grounded at any point as this will bypass the squelch circuit and make it inoperative.

Tom Clerk, WA2YUD

85 cents

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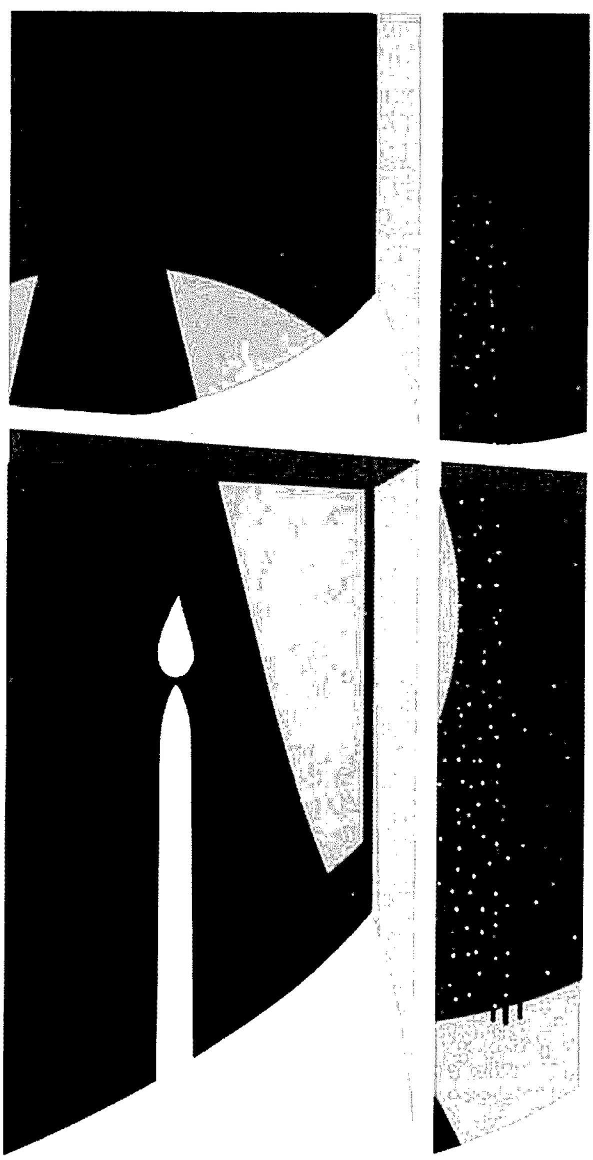
ham radio magazine

DECEMBER 1974



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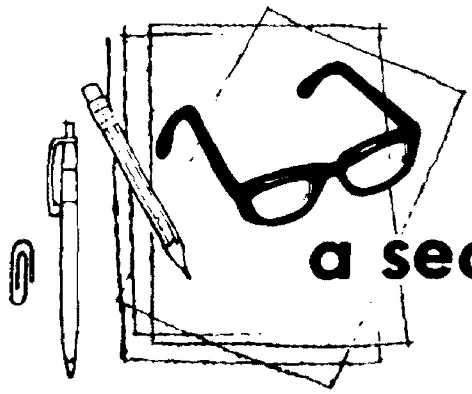


a second look

by jim fisk

Earlier this fall, an engineer from General Electric made news, if not the headlines, with a five-watt handie-talkie and an antenna fashioned from an old golf umbrella. His stunt? He had sent a Morse message through a communications satellite by depressing the mike switch on his handie-talkie, dramatically demonstrating the potential of space satellites for earthbound search-and-rescue missions. I don't want to belittle his accomplishment, but many amateurs whose two-meter equipment was limited to small, hand-held fm transceivers used almost exactly the same technique to work through Oscar 6 during its early orbits.

Nevertheless, the long-distance transmission showed that simple radio gear and a collapsible antenna – plus a space satellite orbiting somewhere overhead – would enable persons in distress to summon help from any point on earth. In the demonstration, conducted for officials of NASA, Roy Anderson sent the Morse message from NASA Headquarters in Washington to the ATS-3 satellite in geostationary orbit over the mouth of the Amazon River, which in turn relayed the signals to GE's Radio-Optical Observatory near Schenectady, New York, which is equipped with a 30-foot dish.

After receiving the message, Observatory personnel transmitted voice signals back through the satellite to Anderson, showing that downed pilots, survivors of shipwrecks and others in need of help could readily receive a voice reply from a search-and-rescue station, acknowledging the SOS and providing rescue information.

Anderson, who holds the basic patent on locating vehicles by ranging measurements from satellites, has proposed a global search-and-rescue system which would require only six geostationary satellites to provide worldwide coverage (except for the polar regions). The satellites would be monitored by three ground stations that could use ranging measurements to locate persons in trouble, and then dispatch assistance. The six satellites could routinely be used for other important activities since the search-and-rescue function would require less than 0.1% of the satellite's transmission power. By equipping future geostationary satellites with a modified antenna, reliable voice signals could even be transmitted from a person in distress to the monitoring station.

On the surface this sounds like a worthwhile proposal, one that could save hundreds, perhaps thousands, of lives. However, who decides what emergency situations should be relayed through the system? Or those that should not? Downed airplanes, foundering ships and other major disasters obviously qualify. How about a man and his family whose Jeep breaks down in the desert? Or a hunter lost in the mountains? The system couldn't possibly handle all the emergencies that occur on the earth at any one time, but human nature being what it is, if everyone is allowed access to the system, it would shortly be hopelessly clogged. On the other hand, not allowing everyone access to the search-and-rescue system defeats its whole purpose. Who is to make that possibly life or death decision?

Jim Fisk, W1DTY
editor-in-chief

presstop

OSCAR 7 LAUNCH AGAIN DELAYED, due to continuing problems with Thor-Delta launch vehicle, but should be in orbit by the time you receive this issue. The new satellite won't be available for use for several days after it is finally launched, as it has to stabilize and undergo complex telemetry checkout -- after that, however, its round-the-clock operation will greatly increase satellite communicators' QSO time.

AMSAT Has Some Fine Slides of Oscar 7, other material for a first-class club program, available on request. Write to AMSAT, Box 27, Washington, D.C. 20044.

HR Report will become the "official information source" for AMSAT news in an arrangement currently being worked out. Details will be announced shortly by AMSAT.

FCC COMMISSIONER LEE APPLAUDS AMATEUR SERVICE in "A Tribute and a Challenge," his banquet speech before the Quarter Century Wireless Association annual convention in Orlando, Florida. Lee's speech was both a review and a look forward. He had obviously done his homework, particularly with some of the people who are preparing for the 1979 World Administrative Conference of the ITU. He endorsed the proposals to eliminate sharing of the amateur bands with other services and to establish new bands at 10.1, 18.1 and 24 MHz, recognizing that a great deal of effort would be required before adequate support can be mustered for such far-reaching ideas to have any chance of success.

REPEATER AUTOMATIC TRANSMISSION of "public service" information is the subject of a recent FCC policy letter. The problem is that unrequested regular weather or time announcements can be interpreted as "broadcasting" and are thus illegal; any repeater with automatic time ID, weather or similar automatically transmitted information must discontinue it as soon as possible.

Repeaters May Still Provide this type information, but only on request. It's OK to dial up a phone company's time or weather with an autopatch, or program a repeater to transmit time and ID for logging purposes while it's being used.

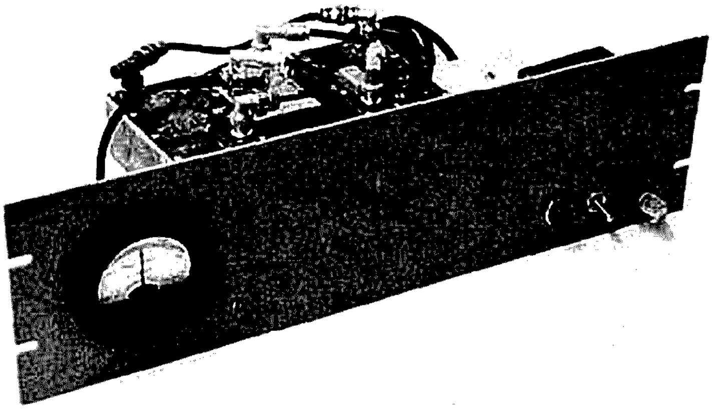
FCC "ORIGINAL AND 14 COPIES" REQUIREMENT attacked in Petition for Rule Making submitted by WB5BBH and WA5VTA on behalf of the Handicappers Information Net. Though there are good logistical reasons for continuance of this FCC requirement, there is no doubt that it stifles amateur participation in the rule-making process.

Ham Radio Magazine will help anyone without access to a copy machine meet the 14 copy requirement. Simply send your original response to Ham Radio, Greenville, NH 03048 and we'll make and collate 14 copies and send them First Class mail to the FCC, all for a less-than-cost $1.00 per original page.

FCC LICENSE BACKLOG is on the increase again, after having been under control just a few months ago. Turn-around time for a "typical" amateur application is running about 25 days with any special considerations likely to extend those times considerably. FCC is working very hard to reverse the situation, has even started contracting for outside data-handling assistance.

RSGB PLANNING HISTORICAL UPDATE, and editor Ron Ham is looking for useful inputs particularly concerning the last ten years. RSGB's first half century (1913-1963) was covered in "World At Their Fingertips," and now the RSGB plans a sequel. Anyone with info to contribute should write to Ron Ham, Faraday, Greyfriers, Storrington, Sussex, England as soon as possible.

SEVERE RECESSION HITS JAPANESE ELECTRONICS INDUSTRY, with a number of firms reported to be threatened with failure. Parts industry is operating at an estimated 60% of capacity, with lay-offs and early retirements widespread. No firms directly involved with the amateur field have as yet been reported in difficulty, and it may well be the U.S. and European amateur markets that are keeping those firms healthy.



frequency synthesizer

for 220 MHz

Construction details for a simple frequency synthesizer for the 220-MHz amateur band

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

The business of generating a crystal-controlled 220-MHz signal is usually accomplished by using a high-frequency crystal oscillator followed by frequency multipliers. For example, you can use any of the systems described in fig. 2. The multiplier chain using an 8-MHz crystal (fig. 2A) has a familiar look because it is much like many two-meter systems that use surplus FT243 crystals. In fact, crystals between 8.148 and 8.222 MHz may be used for *both* 2 meters and 220 MHz (8150-, 8175- and 8200-kHz crystals in the standard FT243 series).

At about 20 MHz, fundamental-mode crystals are replaced (in availability) by higher-mode types. This means that the systems in figs. 2A through 2D will probably use fundamental-mode crystals, in parallel-resonance, to control their oscillators. The systems in figs. 2E through 2G use higher-mode crystals (3rd or 5th overtone) in series-resonance. As it happens, there are some surplus crystal types (CR-8/U and CR-24/U) that encompass this latter 18.333- to 27.5-MHz region.

By purchasing more expensive higher-mode crystals which operate in the vhf

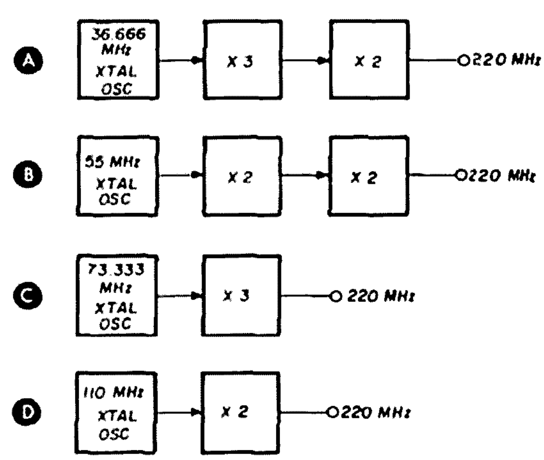


fig. 1. Frequency multipliers that use higher-mode crystals require less multiplier stages and provide greater separation between undesired harmonics.

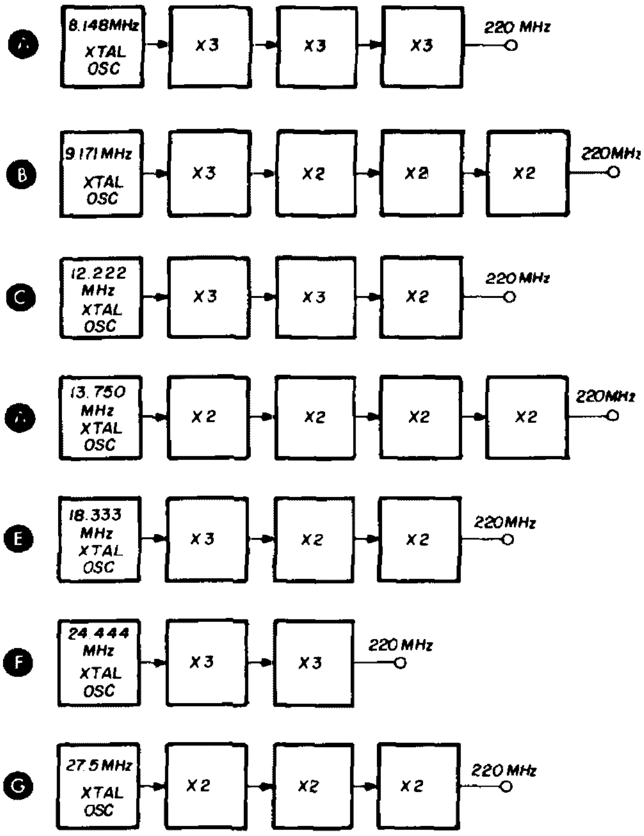


fig. 2. Several frequency-multiplying schemes using low-frequency crystals which are suitable for operation on 220 MHz.

spectrum, the number of multipliers can be reduced. Several systems of this type are shown in **fig. 1**. The use of vhf crystals has one important advantage: the Q of the tuned circuits in the frequency multipliers becomes less important because of the wider *percentage* difference between undesired, adjacent harmonics. That is, when you multiply 8.148 MHz by 27 to get to 220 MHz, there is also a probability of producing times-26 (211.8 MHz) and times-28 (228.1 MHz). If significant 211.8-MHz energy is present, you will no doubt be hearing from your neighbors trying to see the football game on Channel 12. On the other hand, multiplying to 220 MHz from 110 MHz involves only one step: times 2. While it's possible to have both 110-MHz and 330-MHz energy present in the output (*times 1 and times 3*), these other components differ by such large frequency *percentages* the output tank circuit usually discriminates against them.

There is also a difficulty which arises when using vhf crystal oscillators and few multiplier stages. This difficulty is in the crystal oscillator itself, and is in addition to the fact that the crystals themselves are not usually available at low cost. The vhf crystals used in the systems shown in **fig. 1** are usually 5th-, 7th- or 9th- overtone types. To insure that the crystal oscillates on the desired mode, the oscillator circuit must have some built-in mode suppression to prevent oscillation at lower modes. The oscillator has a tendency to oscillate in lower modes simply because the Q of the crystal is generally higher in these *lower* modes (i.e., the series resistance at series resonance is lower).

For third- and fifth-overtone crystals a simple parallel circuit, resonant at the desired frequency of operation, formed by the crystal holder capacitance and an added inductor is often adequate. Such a circuit is shown in **fig. 3**. For 7th- and 9th-overtone oscillators it is sometimes necessary to add series-resonant traps at the frequencies of the undesired lower-frequency modes as shown in **fig. 4**. In short, 7th- and 9th-overtone crystals tend to have more complex oscillator circuits, requiring more critical tuning.

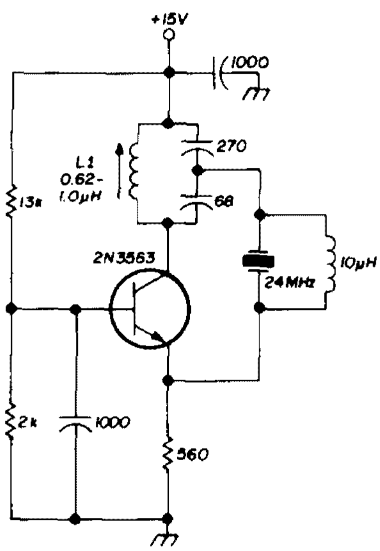


fig. 3. Crystal oscillator for third- and fifth-overtone crystals. Mode suppression is provided by the 10-μH inductor which, with the 4.5-pF holder capacitance of the crystal, is series resonant at 24 MHz.

A relatively new method of frequency multiplication that has become practical with modern frequency synthesis techniques is shown in fig. 5. With the newer ECL (emitter-coupled logic) ICs capable of frequency division up to 1 GHz, the 220-MHz band falls easily within the synthesis method of frequency multiplication. The particular system shown in fig. 5 uses a divide-by-ten from a 220-MHz vco and a 22-MHz crystal oscillator, but the choice of multiplication ratio is almost arbitrary with this method. The primary requirement for being able to multiply by N is that you find a way to digitally divide by N. Since, at least at lower frequencies, N can be any integer, frequency multiplication by even large prime numbers is possible.

For an actual circuit look at the times-10 multiplier shown in fig. 6. Here, a Fairchild 95H90 (U3) is used to divide the vco output frequency by 10. The 22-MHz output of the 95H90 is compared in phase with the output of the 22-MHz crystal-controlled oscillator. The phase comparator (U1) in fig. 6 is a

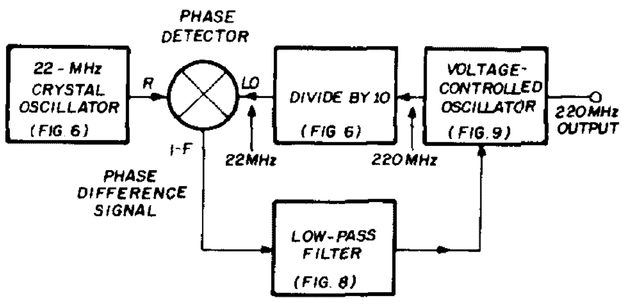
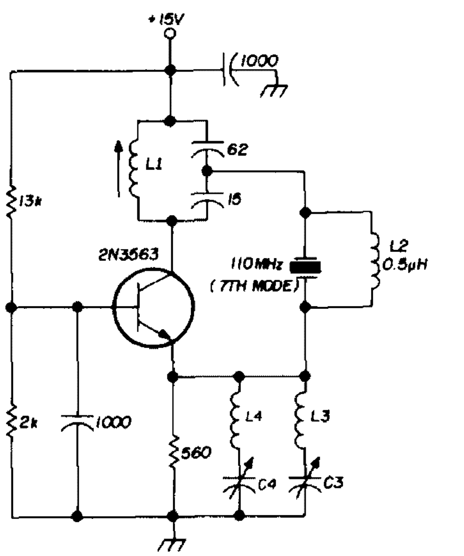


fig. 5. Basic phase-locked frequency synthesizer for 220 MHz. In this circuit a doubly-balanced mixer is used as a phase detector.



C3-L3 series resonant at approximately 5/7(110 MHz), 1 μH and 0.6-10 pF

C4-L4 series resonant at approximately 3/7(110 MHz), 2.2 μH and 0.6-10 pF

L2 0.5 μH (parallel resonant at 110 MHz with 4.5-pF holder capacitance)

fig. 4. In crystal oscillators designed for 7th- and 9th-overtone crystals it is often necessary to include series-resonant traps at the frequencies of the lower, undesired modes. In this circuit L3-C3 are resonant at 5/7 the output frequency and L4-C4 are resonant at 3/7 the output frequency. Inductor L2 in parallel with the crystal holder capacitance is resonant at the desired output frequency.

standard doubly-balanced mixer, manufactured by a number of firms. The doubly-balanced mixer can even be home made,[1] but there are several commercial units available for less than $10. **Fig. 7** shows how a typical doubly-balanced-mixer performs as a phase detector; note that the dc output voltage is only a few tenths of a volt.

The output of the phase detector is passed through a special form of active low-pass filter. This low-pass filter (**fig. 8**) is often called a tracking filter, and one form of it or another is almost always used in phase-locked-loop systems. The active low-pass filter shown here not only provides filtering, it also provides gain. The gain makes up for the low sensitivity of the doubly-balanced mixer used as the phase detector. The maximum gain of the active filter (at dc) is ten, the ratio of 1 megohm to 100 kilohms.

The output of the tracking filter is used to control the vco. The vco circuit and its buffer amplifier are shown in fig. 9. The vco is a type of Colpitts oscillator commonly used at vhf, modified to allow for voltage-tuning by means of diodes CR1 and CR2 – 6.8-pF varicaps (capacitance specified at a reverse bias of four volts). In this circuit they are operated in series with a reverse bias of five volts. Therefore, the total capacitance across

inductor L3 from varicaps CR1 and CR2 is only about 3 pF.

A very simple buffer amplifier using a common gate fet (Q3) is used to isolate the vco from the stages it drives. The gate buffer are in one cast-aluminum box Pomona 2906). The tracking filter is in a small aluminum box (LBM-00) and the rest of the rf circuitry is in a second cast-aluminum box (Pomona 2906).

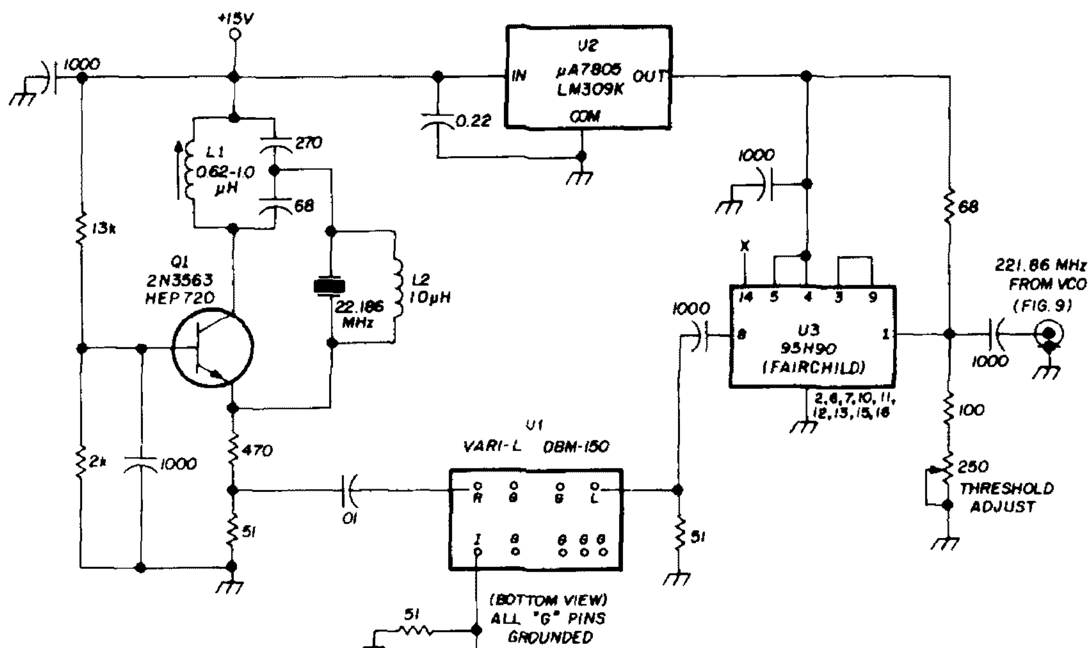


fig. 6. Crystal oscillator (Q1), phase detector (U1) and divide-by-10 counter (U3). U2 is a 5-volt voltage regulator for the 95H90. Phase detector output vs phase difference is plotted in fig. 7.

is at dc ground and the source is untuned; the dc source current flows through the link on L3. The output link on L5 is used to couple the 220-MHz signal back to the input of U3 (fig. 6). A second link may be used to couple 220-MHz energy out to succeeding amplifier stages, but I only used one – mismatching a bit.

A dual regulated power supply was used to provide the plus and minus 15-volt supplies needed for the operational amplifiers (U4 and U5); +5 volts is derived from the +15 volt line using two three-terminal voltage regulators (U2 and U6). Separate five-volt regulators were used to power U3 and vco (Q2) because of possible coupling through the power supplies. The power supply is shown in fig. 10.

construction

The photograph shows the complete 220-MHz system. Note that the vco and Three short coaxial cables connect the three enclosures. It is important to tie the crystal down with a copper strap, as shown, for grounding and acoustical reasons.

The rf circuitry in the two cast-

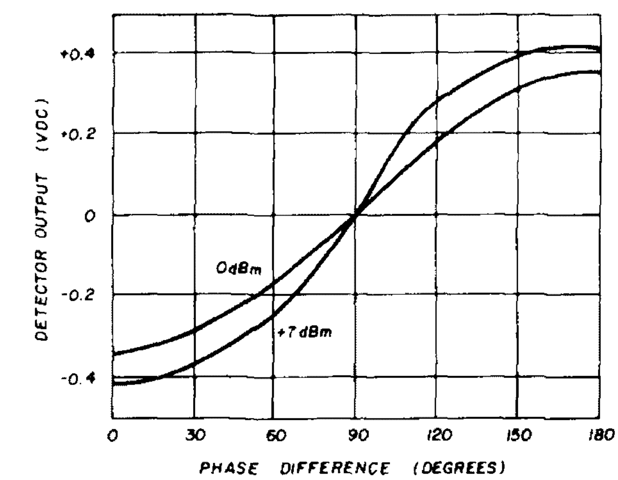
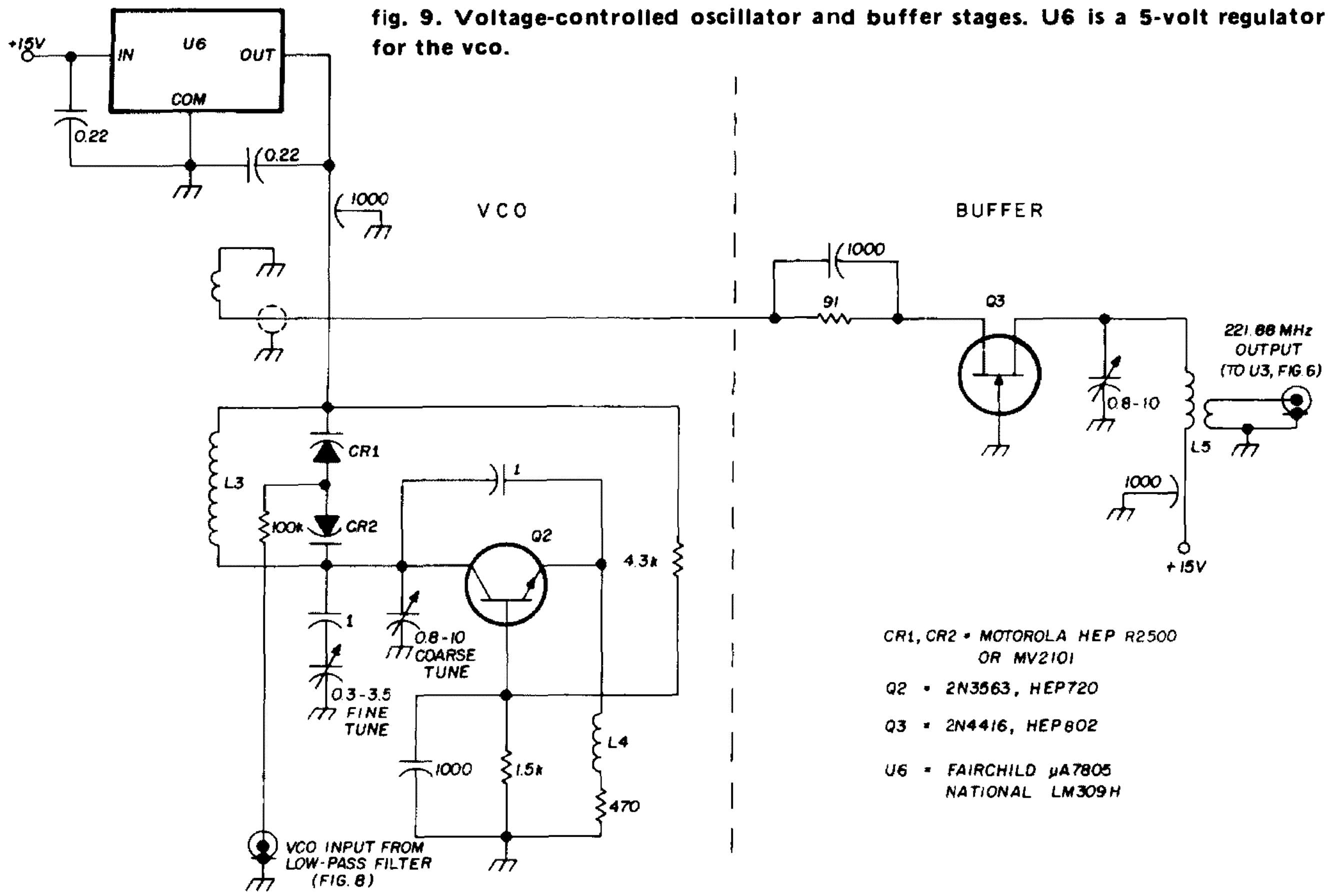


fig. 7. Typical dc output vs phase difference for the doubly-balanced mixer phase detector (U1 in fig. 6).

fig. 9. Voltage-controlled oscillator and buffer stages. U6 is a 5-volt regulator for the vco.



L3 3½ turns no. 20, 3/16" (4.5 mm) ID, 1/4" (6.5 mm) long. Output link is 1 turn on cold end of L3

L4 5 turns no. 28, 1/8" (3 mm) ID, 3/16" (4.5 mm) long

L5 hairpin of no. 16 wire, 1.5" (76 mm) long, 7/16" (11 mm) wide, spaced 1/4" (6.5 mm) above board. Output link is 1" (25.5 mm) long

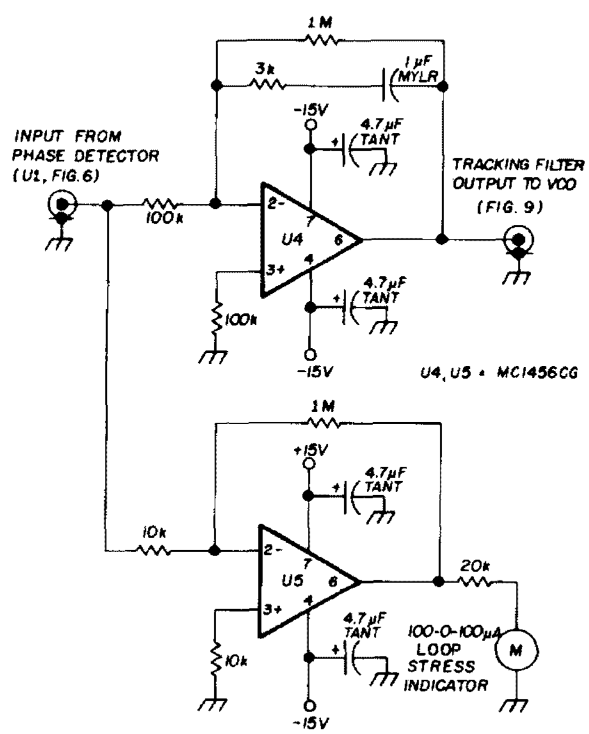


fig. 8. Active low-pass filter used in the 220-MHz frequency synthesizer. U4 and U5 are operational amplifier ICs.

aluminum boxes was actually built on pieces of copper-clad PC board the same size as the plates which come with the boxes. This allows for ease of soldering grounds and provides a near perfect copper ground plane.

alignment

Adjustment of the 220-MHz system can be done with a vtvm and a grid-dip meter. It is best to take it one section at a time. First, with the gdo coupled to the crystal oscillator collector coil (L1), tune for maximum oscillation level consistent with reliable starting. The 22-MHz signal at the R port of the doubly-balanced mixer (as measured with an rf probe on the vtvm) should be 0.5 volts rms.

With the vco input shorted to ground (reverse biasing the varicaps with a constant five volts), the vco should be adjusted for oscillation at 220 MHz (using

the gdo as an absorption wavemeter). Then the buffer should be tuned for maximum 220-MHz output.

Connect the vco to the input of U3 (vco input still shorted) and adjust the threshold adjust pot until the divider is triggering. Assuming that the input is still nearly 220 MHz and the output of U3 is nearly 22 MHz, you can repeak all the adjustments except the crystal oscillator and vco frequency. This should result in the output (i-f) port of the doubly-balanced mixer having relatively low-frequency energy present. This can be seen with a scope at the i-f port, or by connecting a meter to the output of op amp U5 and slowly and carefully adjusting the vco frequency. As the vco goes through *exactly* ten times the crystal frequency, the meter will deflect back and forth.

If all goes well, connect the output of the tracking filter to the vco input and re-tweak the vco frequency for a lock. Locking can be observed by a dc reading at the output of U5 which responds directionally to vco tuning. The dc output of the phase-detector (U1) is often called *loop stress* and it is the best indication of the loop being locked or not. For this reason, a meter was added to the amplified phase-detector output (output of U5) for continuous monitoring.

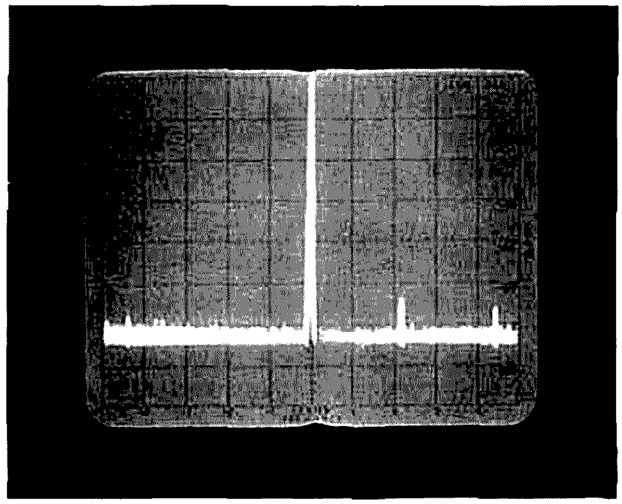


fig. 11. Output spectrum of the 220-MHz frequency synthesizer, measured with a Hewlett-Packard spectrum analyzer. Note the sidebands at ±22 MHz and its multiples — all are more than 50-dB down (horizontal scale is 10 MHz per division).

Several precautions should be mentioned. IC U3 is mounted in a unique way; it is *soldered* in, with the bottom of its ceramic package in contact with the copper laminate and *all* grounded pins soldered down. This is for maximum heat transfer, which *directly* affects the upper frequency at which the IC will count. For more details see references 2 and 3. Do *not* ground pin 14 of U3.

There is also the problem of false

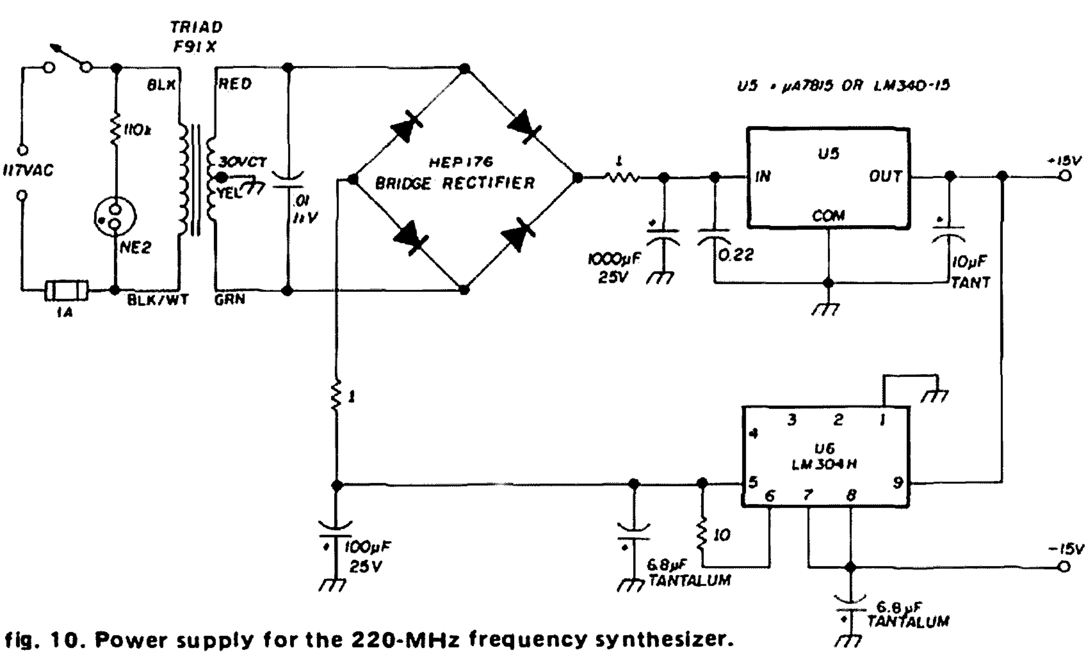


fig. 10. Power supply for the 220-MHz frequency synthesizer.

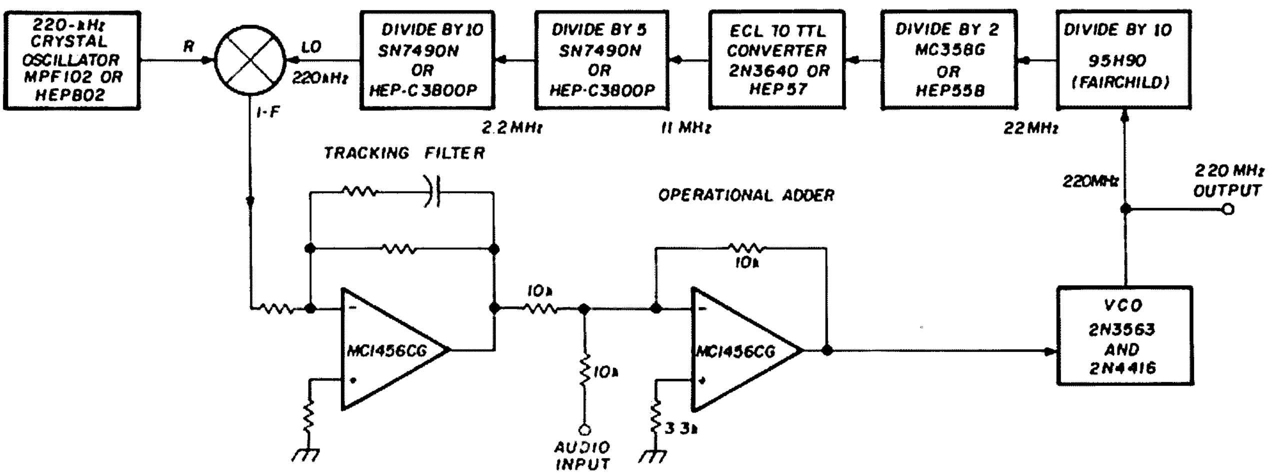
locking; this occurs at the points where the 95H90 is marginally triggering – even with the vco input grounded. It occurs as the vco gets too far from the center frequency of the buffer amplifier's passband and the output begins to fall. Less voltage will cause the counter to miss

carrier *are* visible. but they are all more than 50-dB down.

phase modulation

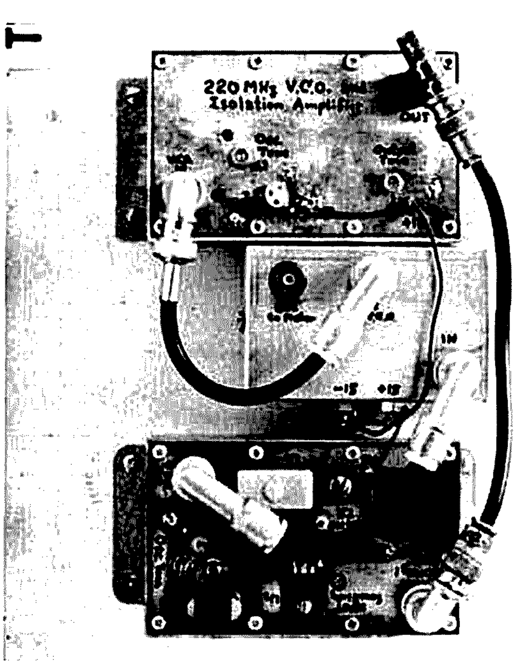
Finally, it must have occured to some of you that it is possible to phase modulate the vco by simply operationally

fig. 12. Basic system for phase modulating a 220-MHz frequency synthesizer. Similar technique could be used for other amateur vhf bands.



counts, and the net affect is a false lock indication.

The final result of proper tuning and care in construction is a signal that looks like that shown in fig. 11. Note that sidebands spaced 22 MHz around the



Circuitry for the 220-MHz frequency synthesizer is packaged in three enclosures which are cabled together.

adding an audio voltage into the vco (after the tracking filter). The only trouble with this is that the audio is only allowed to swing the phase ±90° (at most) as seen at the phase detector. This means the vco phase may be swung ±900° because of the divide-by-ten circuit between it and the phase detector – not much deviation. However, by going to three decades and a 220-kHz crystal, you can get up to ±90,000°. If this seems to be bringing back the days of the old Serrodyne modulation, it is– except that the times-1000 multiplier is easier. Fig. 12 shows a block diagram with suggested digital ICs in a system for phase modulation of this type.

references

1. William Ress, WA6NCT, "Broadband Double-Balanced Modulator," *ham radio*, March, 1970, page 8.
2. Everett Emerson, W6PBC, "Advanced Divide-by-Ten Frequency Scaler," *ham radio*, September, 1972, page 41.
3. Everett Emerson, W6PBC, "Circuit Improvements for the Advanced Frequency Scaler," *ham radio*, October, 1973, page 30.

ham radio

understanding Q

A discussion of the Q of LC tank circuits, and its effect on transmitters, receivers and antennas

Carl C. Drumeller, W5JJ, 5824 N.W. 58 Street, Warr Acres, Oklahoma 73122

In World War I some merchant ships were heavily armed but disguised to appear as unarmed trawlers. German submarines, not wanting to expend an expensive and scarce torpedo on a mere merchant vessel, would surface to sink it with gun fire. Then the Q ship would drop its false sides, revealing mighty guns that would destroy the submarine. Maybe that's how Q got a reputation of being not only a deep mystery but also something not really on the up and up. Many of the references to Q in the literature have done little to dispel the confusion.

Q is a "figure of merit," every textbook tells. If it's good, why don't we use lots of it in the plate tank of a transmitter? Oh, no, say the books; too much Q will make the circuit inefficient! A contradiction? A high Q dissipates little power in the form of heat, but a high-Q antenna is to be avoided like a plague! Why?

A high-Q circuit is one with little resistance, one page of a text tells us; three pages later in the same book you read, "Use a high value of resistance in the circuit so as not to lower its Q." Which do you believe? High Q means a good flywheel effect. Flywheel? What has that got to do with electronics? Q concerns the relation of stored energy to released energy. Oh, so we measure the efficiency of lead-acid storage batteries by their Q? Seemingly, there's just no logic to Q!

There is logic, but you must have a clear

concept of the many faces of Q before that "logic" appears logical! Let's start with a look at **fig. 1**. **Fig. 1**A shows a basic LC circuit, one with provision for applying a momentary pulse of power to it. Although a battery and a switch are shown, the switch could be replaced by a vacuum tube pulsed into conduction by the application of a positive-going spike to its grid. Now glance at **fig. 1**B and imagine a very brief closure of the switch, with its reopening a tiny fraction of a second later. During the small period of time it was closed, energy flowed from the power source into the LC circuit. Because of this ability to accept energy, an LC circuit often is called a tank.

While considering this tank, let's see what the incoming current attempts to do and what it does, step by step. It attempts to flow equally through both legs of the tank. It can't, initially, for the very nature of inductive reaction in the inductance leg retards the effort of the current to traverse that path. But a portion of the current flows unimpeded into the capacitance leg, building up an electrostatic change in the form of excess electrons on the surface of the dielectric adjacent to the upper plate of the capacitor.

If it helps your understanding, you might say that an equal number of electrons flowed out of the dielectric next to the bottom plate, leaving an excess of *holes* there. That might sound more familiar to those of you who have become accustomed to thinking in terms of semiconductors.

While the electrostatic charge was accumulating in the capacitor, the current hadn't abandoned its attempt to flow through the inductor. It was slowly making its way down that leg. As it moved through the inductor, it created an inductive field which spread out from the coil. This field contained, in the form of electromagnetic lines of force, a portion of the initial energy supplied by the power source. As the initial energy pulse was very brief compared to the natural period of the tank (the period = 1/f, where f equals the frequency at which $X_L = X_C$), additional energy from the excess electrons accumulated on the capacitor's upper plate join in the attempt to push current through the inductor. In time, they succeed.

As these incoming electrons neutralize the excess of holes on the lower plate, current flow tries to stop. It can't, just yet. For as it falters, the inductive field collapses. The energy stored in it is returned to the coil, causing a continuing flow of current. But this flow has to stop, too. When it does, the excess electrons built up on the lower plate by the current flow caused by the collapsing magnetic field tries to return to the upper plate, which now has a deficiency of electrons.

This flow of electrons, a current slightly weakened by unavoidable losses, meets the same obstacles as did the initial flow through the inductor. Like the other it succeeds, and one cycle of oscillation has taken place in the LC tank. Then starts a new cycle of oscillation, duplicating the original cycle but less the pulse of energy from the external power source. Again the cycle is accomplished. But, this time, the magnitude of current flowing, the magnitude of electrostatic potential built up on the capacitor, and the magnitude of electromagnetic force built up in and returned from the inductor's field will all be less than previous cycles. This is where Q comes in.

circuit losses

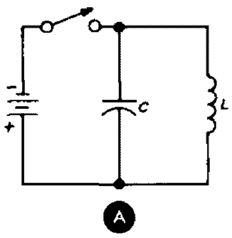
Let's think about *why* those three magnitudes decayed. The decay was caused by losses. Where did the losses lurk? The capacitor leg is a good place to start. This is an imperfect world, and no insulator is perfect; therefore some losses came about from leakage across the insulation incidental to the capacitor. Even though the capacitor may have had silver plates, some ohmic losses were present. And the dielectric itself contributes to the loss total by requiring the expenditure of

some energy to rearrange its molecular structure in order to accommodate the excess of electrons (or holes) on first one plate and then the other. All of these losses are additive.

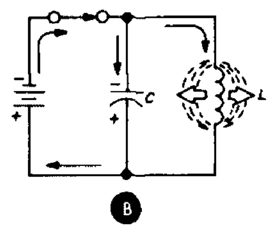
In the low-frequency (30 to 300 kHz), medium-frequency (300 kHz to 3 MHz) and high-frequency (3 to 30 MHz) ranges, the total losses associated with the capacitor are so low they usually are not considered. The capacitor accepts energy during one half-cycle and restores it to circulation during the following half-cycle with an efficiency approaching perfection!

Instead, for those frequencies, we look to the inductor to find the main source of loss. Like the capacitor, it has certain insulation losses, divided between leakage and dielectric hysteresis. Unique to it, however, is the fact that not all the electromagnetic force stored in its field is restored to the inductor when the field collapses. Some of it is radiated, some of it is transliterated into heat by hysteresis effect in nearby metallic and dielectric objects. All of these add up to a quite considerable sum of losses. So large, in fact, that we think of the inductor exclusively when we talk about losses that deteriorate Q in a high-frequency circuit.

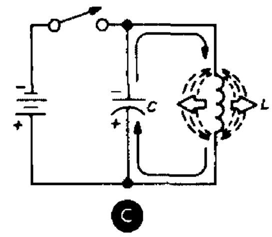
You should not lose sight of the phenomenon of energy being extracted from the tank and stored in the electromagnetic field during one part of the oscillation cycle and then being returned to the inductor during the next half-cycle. If the inductor losses are low, then a very substantial portion of the stored power will be recaptured. Low losses contribute to a high circuit Q so we associate high Q with a high percentage of



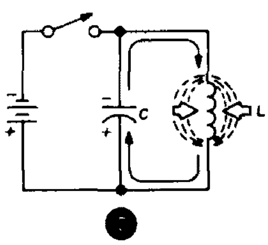
A. Initial state.



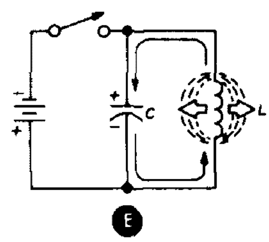
B. Start of initial charge half-cycle. Electromagnetic force being stored in field of inductor and electrostatic charge being stored in capacitor.



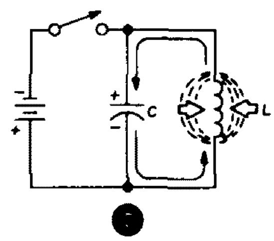
C. Continuation of first half-cycle. Electrostatic charge contributing to build up of electromagnetic force.



D. Final phase of first half-cycle. Electromagnetic field collapsing and returning energy to the capacitor.



E. Start of second half-cycle. Electrostatic charge causes flow through inductor, buildup of electromagnetic field.



F. Final phase of second half-cycle. One full cycle will have been completed. Return to (c) for start of next cycle.

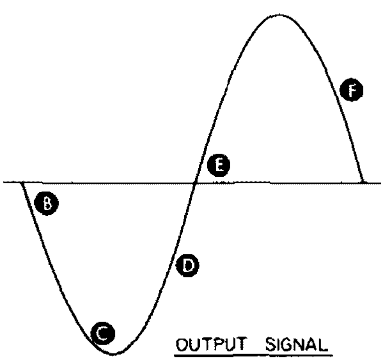


fig. 1. Oscillation cycle in an LC tank circuit.

stored power being returned to the circuit.

In considering the matter of stored and returned power, you should not overlook one striking difference between energy stored in a capacitor and that stored in the field of an inductor. The electrostatic charge stored in a capacitor can be retained there for long periods. A well-insulated capacitor will hold a high percentage of its charge for days. The electromagnetic force contained in the field of an inductor, however, can never be static. It *must* be in a state flux. The moment it ceases expanding from the impetus of current flowing through the coil, then it starts collapsing.

Usually all losses are lumped into one heap when we talk about Q. As these losses dissipate energy as a resistor dissipates energy, it's both convenient and accurate to label these several losses as resistive, to lump them as one, and to refer to the aggregate as the "equivalent resistance" of the circuit. That agreed upon, let's look at **fig. 2**, which shows circuits with lumped losses depicted as resistors. **Fig. 2A** has the resistor in series with the inductive leg. In such a circuit, losses are lower when the resistance is low. With losses low, Q is high, and the formula $Q = X/R$ applies. We usually have this circuit in mind when we talk about keeping Q high by cutting resistive losses.

Fig. 2B takes on more meaning when you glance on to **fig. 2C**. You know from experience that having a grid resistor that is too low in value results in circuit losses that reduce both signal strength and circuit selectivity. For this arrangement you'd use the Q formula $Q = R/X$.

With a little mathematical juggling, you can transform the circuit in **fig. 2A** to that of **2B** or vice versa. As this article is concerned only with identifying Q, I'll refer you to any of the numerous texts that explain the mathematical manipulations.

Knowing that high Q relates to low circuit losses, let's talk about ways of increasing Q by decreasing effective resistance. We'll limit our consideration to circuits in the low-, medium- and high-frequency spectrum. That means we'll be talking about only the inductive leg.

practical inductors

The turns of wire (usually) constituting an inductor provide a fertile ground for Q improvement. If you can have the same inductance with fewer turns (shorter wire length), it stands to reason that ohmic losses will be reduced. This suggests a ferrite core. Good, providing that the proper type of ferrite is used because ferrite is frequency sensitive. Ferrite can be very lossy so be sure that the type you select won't contribute more hysteresis loss than it deletes ohmic loss!

The magnetic field is another good spot for a bit of spade work. The field can be confined by winding the inductor in the form of a toroid. Although air-core toroids have been made (and once were very popular in TRF receivers), most now are made with ferrite or powdered-iron cores. Even better than the toroid is the cup-core or pot-core configuration. Although very effective in field containment, it's not convenient to work with, is rather expensive, and is not often used by amateurs.

If you prefer not using a confined-field type of inductor, you must be careful not to introduce excessive loss by mounting the inductor too close to other objects. Especially guard against getting it too close to shielding. Copper is bad enough, but iron and steel are much worse! An old rule-of-thumb is to keep the coil at least a half-diameter away from any shielding. Insulators also introduce loss so keep down the amount of insulating material in the inductor's field. Air-core coils have quite small losses, especially when wound with spaced turns, and ridged coil forms have lower losses than those which provide continuous support to the wire. Since some insulating materials have much lower loss than others, investigate and select the type that'll serve you best.

Thus far we've talked about the Q of the tank circuit by itself, but tank circuits just don't live that way in real life. You'll always find them associated with other

circuits or circuit elements. These associations inevitably tend to reduce the Q. This is too bad, for you make careful effort to keep the tank's Q high, then, when you put it to use, the Q is sliced down in a disheartening manner. However, don't let this situation keep you from designing and using a high-Q tank because any losses caused by a low Q in the basic tank are lost to you forever! On the other hand, the lowered Q that comes about from coupling the tank to other circuits may mean only that you've used

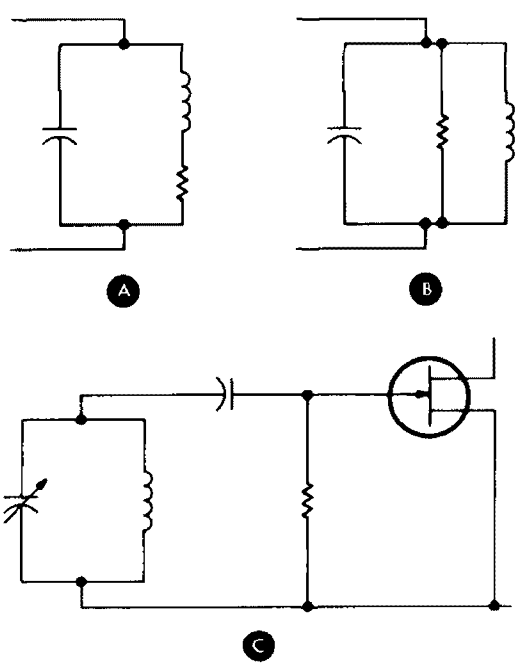


fig. 2. Resistance in practical LC circuits. Series resistance in (A) and parallel resistance in (B) are contributed by the LC components used in the circuit. Resistance in (C) is gate resistor of following fet stage.

power from the tank, used to excite a following stage or to be radiated from an antenna. So it was not lost, just transformed.

Let's look at a circuit that lowers Q yet serves a desired purpose (fig.3). This is the plate tank circuit of a transmitter. It's coupled, by means of an adjustable pick-up link, to an antenna through a 72-ohm transmission line. We'll assume that the line is matched to the feedpoint of a resonant antenna so there'll be a 72-ohm resistive load presented to the pick-up link. This is a form of output coupling that was in common use 30 or 35 years ago.

With the link very loosely coupled to the tank coil, very little of the 72-ohm load will be reflected into the inductive leg of the tank. The tank's impedance (and Q), therefore, will be high. When the tank is tuned to resonance only a little plate current will flow. When tuning into resonance, a sharp and deep dip in plate current will be seen. When the link is moved into closer relationship with the plate coil it will reflect more resistance into the tank and its impedance (and Q) will decrease. Plate current will increase. The plate current dip, at resonance, will be broad and shallow. Power is being extracted from the tank and fed to the antenna. You don't regret lowered Q in such instances!

This circuit depicts another contributor to lowered Q. It's the vacuum tube supplying power to the tank. Every power-generating (or power-converting) device has internal resistance. The vacuum tube is no exception. Its resistance is in parallel with the tank, as shown in fig. 2B. The Q of the tank circuit, therefore, is lowered.

flywheel effects

While looking at the circuit of fig. 3 let's think about another aspect of Q, the flywheel effect. The vacuum tube, unless it's operating class A, does not supply a steady flow of power into the plate tank. Instead, the power is applied in pulses. As the tank is tuned (synchronized) to the frequency (repetition rate) of the pulses, a burst of power is fed to the tank in such a time relation that it is in phase coincidence with the power circulating within the tank. To illustrate, at the moment when the top plate of the capacitor is providing its excess electrons to reinforce those pushing their way down through the inductor, the plate of the tube also is providing a pulse of current to further reinforce the flow.

You'll recall from the foregoing discussion that, unless "recharged" from an external power source, each reversal of

the oscillating electron flow in the tank circuit results in less current and less voltage than the preceding one. Nevertheless, there is a current flow and there is a voltage developed. This phenomenon, the fact that current continues to flow after the initiating energy pulse has been cut off, is known as the flywheel effect. It is desirable that there be a very minimum of deterioration of power circulating in the tank between energy pulses because any drop in power is conducive to the generation of harmonics.

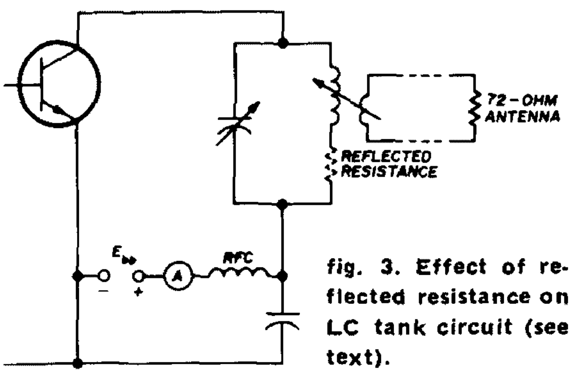


fig. 3. Effect of reflected resistance on LC tank circuit (see text).

Also touched upon previously is the fact that low circuit losses tend to keep the circulating current constant. This leads to the conclusion that low losses indicate high Q which, in turn, means good flywheel effect.

loaded and unloaded Q

Now, let's take up the matter of unloaded and loaded Q. It's not complex. The tank circuit, isolated from all else exhibits unloaded Q. When you associate it with anything else, the Q will deteriorate; this is loaded Q. Usually loaded Q refers to the Q at some stipulated load and the total load is often made up of several contributory loads.

Thus far I've talked about tank circuits in relation to transmitters. There's a reason for this. With the meters associated with a well-designed transmitter you can observe the effects of changes in Q and the manipulations that cause changes in Q. This is not so easily done in receivers.

Before leaving transmitters, let's consider the importance of Q. The prime purpose of a transmitter is to produce a signal on one selected frequency. It is not desired to produce signals on harmonics of that frequency or upon any other spurious frequency. It's unfortunately true that all efficient generators of radio-frequency power tend to generate something other than a pure sinusoidal wave; they generate waves rich in harmonics. A high-Q tank circuit introduces a healthy element of selectivity into the situation. The tank selects the desired frequency, passes it, and rejects (to a degree) all others including those troublesome harmonics. So, following an active device (vacuum tube, transistor, etc.) in an rf circuit we like to insert a tank circuit of moderate Q.

Why "moderate" Q? Let's go back to fig. 1. You'll recall that, as this is a resonant circuit, $X_L = X_C$, and therefore if L is reduced in an effort to raise Q by reducing the length of wire in the inductor, C must be made larger to restore resonance. Circuit power remains unchanged. To accommodate this power, a tremendous store of electrons must accumulate on one plate of the capacitor. As the circuit oscillates, this great store of electrons must flow through the inductor to reach the other plate, creating a much heavier current flow than would have been the case had the capacitance been less and the inductance greater. The heavy current encounters some ohmic resistance in the coil, which results in the generation of heat. For heat, read, "unretrievable loss of radio-frequency power." Not only is power lost, but the resultant heat often damages the coil and adjacent components. So, you see, the effort to increase efficiency by going too enthusiastically after high Q can lead to greatly reduced efficiency. Here, as in many other aspects of life, moderation is the keyword!

receiver selectivity

In association with receivers, Q performs perhaps an even more important role than in transmitters. Although the

growing use of filters (active, crystal, mechanical and ceramic) for setting the ultimate selectivity of a receiver has taken over a function in fixed-frequency circuits that was once reserved for high-Q LC tanks, there are applications for which no better alternatives have been found.

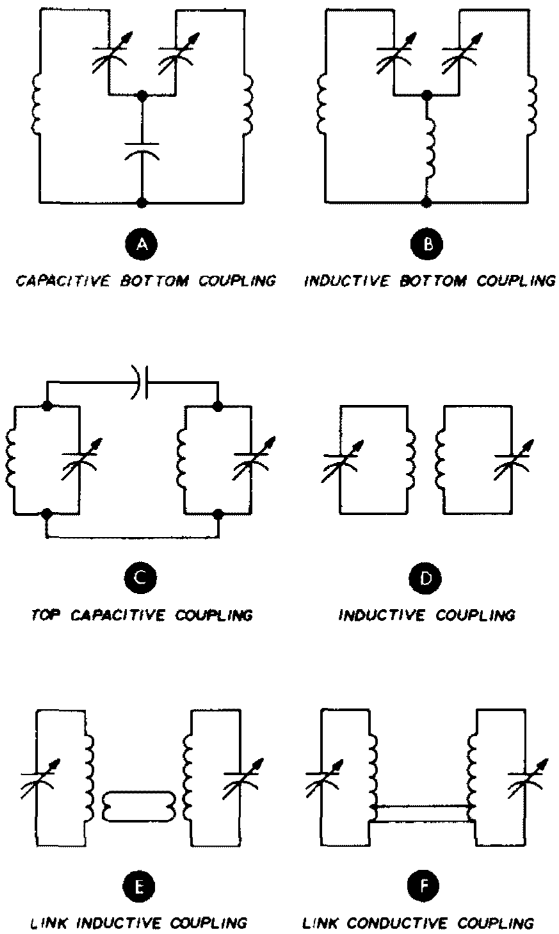


fig. 4. Common methods of coupling LC tank circuits.

These are in the tunable section of a receiver.

Whether a receiver's circuit is the classic TRF, the conventional superheterodyne with a fixed intermediate frequency, or a superheterodyne with semi-tuned input and variable i-f, current practice involves the use of LC circuits to establish a certain degree of selectivity.

Ideally, a receiver should have all needed selectivity before the first active device. Whether that active device is a vacuum tube or a transistor, it deteriorates the performance of the receiver. Why? Because, with the possible exception of the beam-deflection tube, it introduces a degree of non-linearity. Non-linearity means intermodulation products can be generated when a strong undesired signal is present along with the desired signal. Once generated, these products are very difficult to cope with. So you'd like to eliminate that strong undesired signal before it reaches an active device. This calls for highly-selective tuned circuits. One way of getting these is by using LC circuits of sufficiently-high Q.

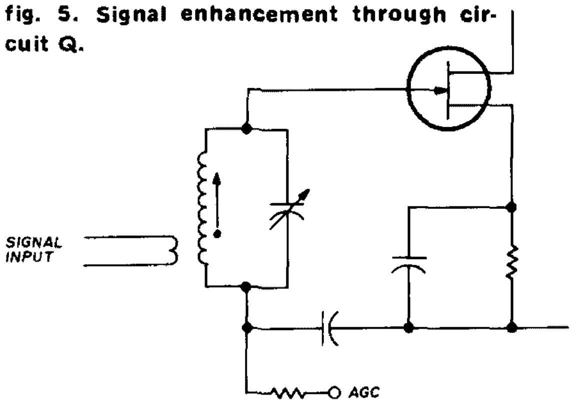
Only by the use of superconductivity can a single LC tank achieve such a remarkable Q, but cryogenic superconductivity is expensive so its use is limited almost exclusively to receivers for reception of signals from outer space. The designers of ordinary receivers must look to other means of achieving selectivity. Fortunately, a ready solution lies in the fact that the Q of two or more cascaded circuits are multiplicative. For example, if you have two cascaded tuned LC circuits, each with a Q of 10, the total Q of the chain is 100. Add another like circuit, and the total Q becomes 1000. This phenomenon permits achieving the high Q needed for reasonable selectivity but does so at the cost of requiring a multiplicity of tuned circuits, circuits that must be ganged and tracked for convenience in tuning. Each of these tuned circuits introduces some unavoidable losses, but it's a price you must pay.

There are many ways of cascading tuned circuits. Each has its proponents, but there seems to be little difference in their performance. Several of the many available circuits are shown in **fig. 4**. Two bottom-coupled circuits appear in **4A** and **4B**, and **4C** shows top-coupling. Conventional inductive coupling is illustrated in **4D**. Link coupling appears in **4E**, with a variant in **4F**. The choice of which circuit to use seems to lie with consideration of physical rather than electrical characteristics.

The magnitude of Q needed to achieve a specific amount of selectivity, say, 6-dB down at 10-kHz bandwidth, varies

with the frequency of the signal being processed. When bandwidth is an appreciable fraction of the signal's frequency, selectivity can be had with reasonable Q. On the other hand, if the bandwidth is very small in relation to the frequency, extremely high values of Q are needed.

fig. 5. Signal enhancement through circuit Q.



Other than the matter of selectivity, there's another aspect of Q that's important to the receiver designer. It concerns the voltage presented to a vacuum tube's grid or an fet's gate. Look at fig. 5. If a voltage *e* is induced from the adjacent link into the inductive leg of the tank, the magnitude of the voltage available between the grid and cathode of the tube will be *e* times Q. Therefore, high Q in receiver tank circuits contributes to the overall gain of the receiver.

antenna Q

When we consider Q in relation to antennas, several factors must be kept in mind. Usually, the ohmic resistance (and, therefore, the ohmic loss) is so small it is swallowed by the much greater "radiation resistance" of the antenna. Quotes are used to head off any assumption that the term has anything to do with real resistance. An antenna, to serve its purpose, radiates radio-frequency power. A resistor will transliterate radio-frequency energy into heat. Each disposes of power so they have a common element of action and the power could be measured in watts by the formula $W = I^2R$. In a resistor true resistance is used in the formula. For the antenna, however, we create an imaginary resistor which, if it existed, would consume the same amount of watts. For example, if a radio-frequency current of two amperes were fed into a 50-ohm resistor, 200 watts of power would be dissipated in the form of heat. If that same two-amperes of current were fed into an antenna and 200 watts of power were fed into an antenna and 200 watts of power were to be radiated into space (less that tiny bit lost in heat because of the small ohmic resistance), then we could conclude that the antenna's "radiation resistance" was 50 ohms.

It would appear that the higher the radiation resistance, the greater (for a given amount of antenna current) would be the radiated power. Unfortunately, that radiation resistance appears as a series resistance in the equivalent circuit of the antenna. What's it going to do to the antenna's Q? Lower it, of course! From these considerations we can conclude that low-Q antennas are desirable, but that's true only when the low Q comes about because of high radiation resistance and not because of high ohmic resistance, high losses or any of the other factors that can lower Q.

summary

In preparing this article, I've intentionally avoided the more conventional approaches for presenting facts relating to Q. There are many excellent texts that deal with such aspects in a thorough and rigorous treatment. I have found *Radio Engineering*, by F.E. Terman[1] especially useful and recommend it highly.

What I've tried to present is an easily-read but factual identification of Q, an account of how it is achieved and enhanced in a circuit, and a limited number of examples of how optimum values of Q are used to accomplish the desired results in transmitters, receivers and antennas.

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1. F. E. Terman, *Electronic and Radio Engineering*, McGraw-Hill, New York, 1952.

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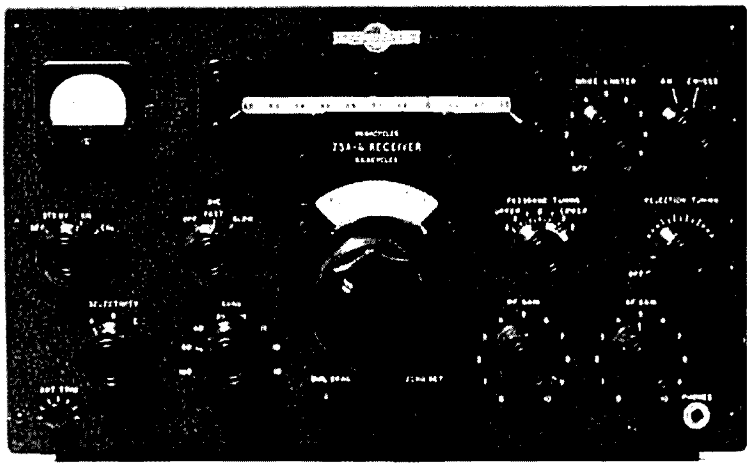


making your Collins 75A4 perform like new

How to clean up the all-too-common tuning problems of an old but popular receiver

Paul D. Rockwell, W3AFM, Kenwood, Chevy Chase, Maryland 20015

Many amateur radio operators believe the Collins 75A4 to be the best amateur receiver ever made. Particularly for CW use, there is much truth to this. Unfortunately, the 75A4 is long out of production and—for some, at least—out of style.

Some of the reasons for this fine receiver's going out of style include: Size, weight and (relatively) high power consumption, old-fashioned appearance (black crackle, square corners), not set up for transceive operation, not equipped for break-in muting, vacuum tube instead of solid-state design, objectionably high noise figure, especially on 10 and 15 meters, and insufficient dynamic range and front-end selectivity.

Of these factors, the latter two are true of *all* receivers, no matter what their vintage, but the 75A4 actually does better with them than almost any current receiver! The noise figure and dynamic range problems have been attacked before,[1] and a good preamp can help the former at the expense of the latter. Another factor, one of the most frustrating and yet most easily overcome, is the age-connected problem of stiff tuning and

frequency jump. Solving this difficulty is the subject of this article.

The 75A4 is at its best as a CW receiver, and CW requires delicate and smooth tuning. As 7A4s age, however, many begin to get stiff and require irregular torque on the tuning knob and some may jump frequency a kHz or two even while not being tuned. Both of these problems have their cause in the permeability-tuned oscillator and dial assemblies. Many amateurs have learned to live with sticky tuning, at least up to a point, but frequency jump is intolerable. It is probably safe to say that these problems account for many of the 75A4s being offered on the market today.

Some discussion of the causes of frequency jump was given in the previously cited article. It is now believed that the two problems are inter-related, and that if sticky tuning is tackled first, the frequency-jump problem will usually disappear along with it.

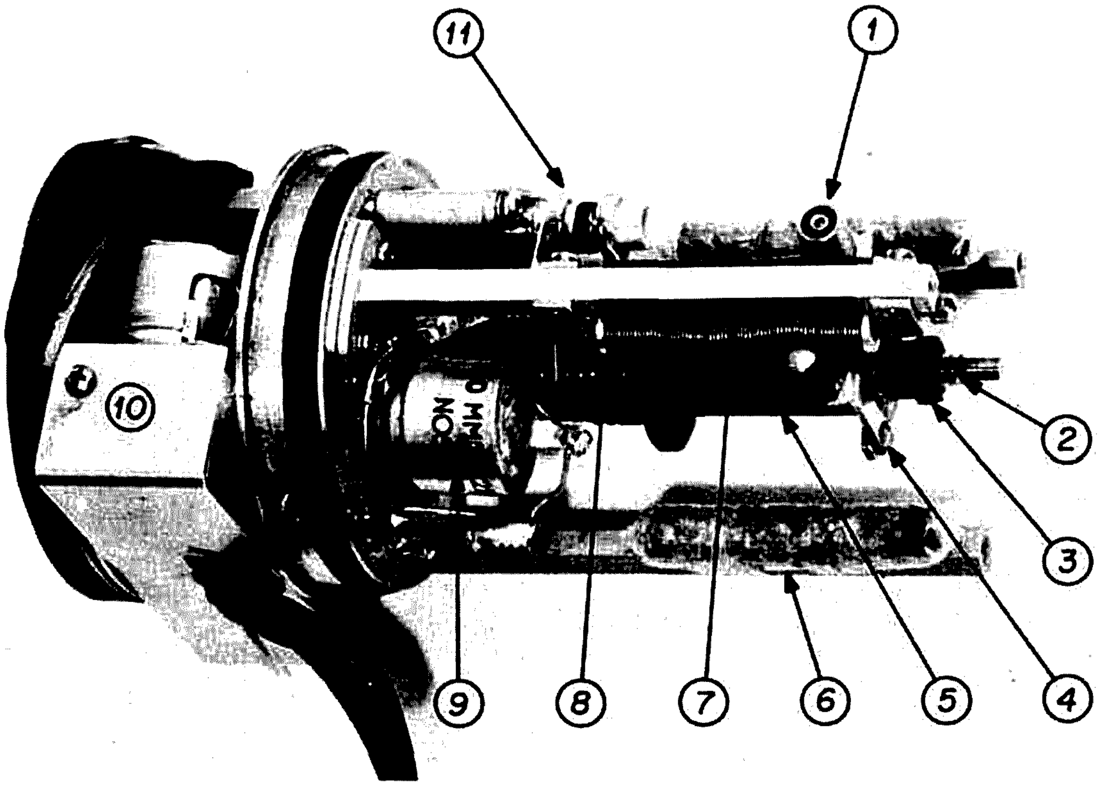


fig. 1. Inside the Collins 75A4 permeability-tuned vfo (PTO). The numbers identify the critical parts of the PTO assembly, described below.

1. Cam idler wheel, which rides on cam stack to control idler.

2. Lead screw, whose rotation moves the tuning slug through the tuning coil.

3. Lead-screw lubricating washer; should be saturated with oil.

4. Cam idler assembly, whose movement makes minor adjustments in oscillator tuning linearity.

5. Tuning slug.

6. Moisture-absorbing silica-gel sacks, blue when dry and pink when saturated.

7. Cam stack, used to compensate for nonlinearities in oscillator tuning.

8. Tuning coil, wound with varying pitch to approximate linear tuning with slug travel.

9. Padding capacitor, which establishes oscillator tuning range.

10. Cover for tube bases and non-critical PTO components.

11. Trimming inductor, used to set the PTO tuning range to precisely one MHz for ten turns of the lead screw.

The bulk of the problem exists inside the PTO (fig. 1), and this is where you are going to have to go. Pay no attention to the manufacturer's caution about not breaking the seal of the PTO—these units were never hermetically sealed, even when brand new. They could breath through the bearings and, perhaps, the rubber O-ring. Moisture-laden air, breathed in a little at a time each time the receiver was turned off and cooled down, usually turned the silica-gel sack pink within the first year's operation—and that was a long time ago. If moisture is a worry, as it might be in a basement shack or in a particularly humid part of the country, you could let the receiver run around the clock (bad from the energy point of view). Better, install a 7½-watt, 115-volt pilot lamp near the PTO, wired directly to the power line, and let it run all the time to keep the PTO warm. At any rate, moisture is not a problem with 99.9% of the 75A4s around, but sticky tuning is present to some degree in almost all of them.

This operation will be a painful one for anyone who doesn't like working with tools. Assuming only the usual number of minor problems along the way, you can expect the complete job of removal, repair and reinstallation of the PTO will consume the better part of a day. If your time and patience are too thin, you might try a partial job—but then don't expect miracles.

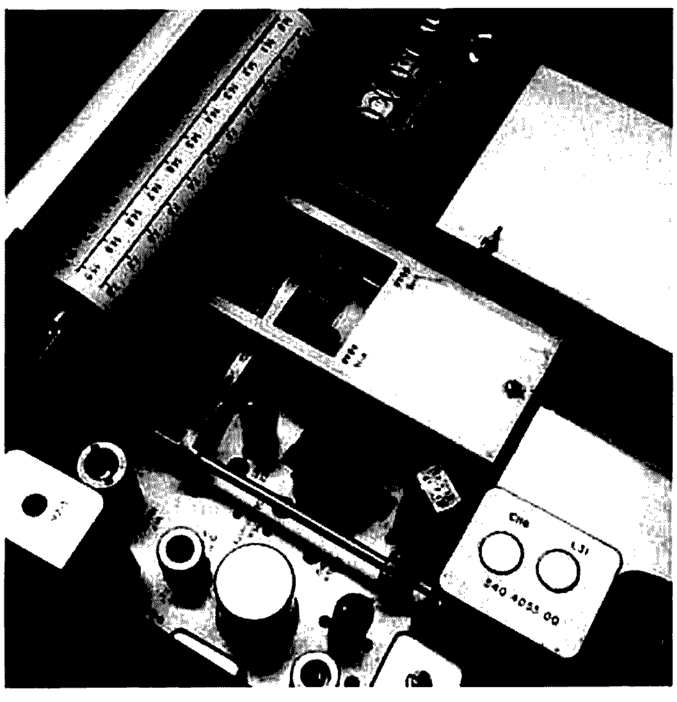
pto removal

The first step (after taking the receiver out of the cabinet) is to set the tuning dial to 14.000 MHz. Next remove the vernier knob, mounting plates, ring gear and pinion. Put the metal parts into a half-pint jar of mineral spirits to soak, or better, clean them in an ultrasonic cleaner if you happen to have one. Make sure all dried grease is removed before you take out the parts and dry them. Set up a row of saucers or ash trays so that all the small hardware can be placed in them in sequence as it is removed. This not only keeps them from getting lost, but is also a nice memory jogger when reassembly time comes.

From the top of the receiver remove the tuner dust cover, top and side screws of the PTO rear cover plate, and set screw and spring of the passband-tuning bronze band. Loosen the two set screws of the tuning shaft, immediately to the rear of the flexible coupling.

Remove the bottom plate from the bottom of the receiver. On the middle-bracing chassis cover plate remove the front two screws and loosen the rear two screws. This will permit the plate to be tilted so the PTO can be pulled out of its shaft coupler when the time comes. Make a sketch of the PTO connections and mark the chassis with a felt-tipped pen to facilitate reconnecting the wiring correctly when the unit is reinstalled. Unsolder the three power leads and the coax.

Pull out the PTO. The first thing to examine and repair is the tuning-shaft grounding wiper. This is the small L-shaped arm at the front, secured by two tiny Phillips-head screws. Its purpose is to provide a good ground return on the shaft



Inside the 75A4. The PTO is hidden by the square cover although the two 6BA6 oscillator tubes, V14 and V15, are clearly visible.

so that rf currents do not have to circulate through the front bearing. When the receiver is new, this wiper rides on the polished finish of the shaft. However, continued use may have caused the shaft to gall at this point—check it with your fingernail. Roughness here can be a major contributor to sticky tuning, so if the shaft is rough loosen the screws, bend the wiper slightly forward so it rides on a smooth portion of the shaft, and apply a touch of grease to the contact point.

lubricants

There are a lot of misconceptions about *lubricants.* For purposes such as this one, plain axle grease and *3-in-1 Oil* are well up on the list. Axle grease is not as strange a choice as it sounds, as pressures (i.e., pounds per square inch) at some contact points can become very high and axle grease is very good at staying put. For those who want something better than axle grease, Aero Shell 7 – a general-purpose aircraft grease—is excellent. However, it is hard to find, expensive, and sells in five-pound (minimum) cans. Shell calls it a "Microgel Diester Synthetic," and it has an operating temperature range of −100 to +300°F. Silicone grease is not good for this purpose because of its inferior high pressure performance.

inside the pto

Now comes the moment of truth! ignoring the red-lettered warning decals, remove the screws holding the PTO cover and carefully slide it off. Examine the PTO assembly, noting the locations of the various components identified in fig. 1. Drop a few drops of *3-in-1 Oil* on the front bearing, on the rear felt washer, and on the cam rollers. Put a dab of grease in the rear sleeve bearing (inside the rear of the PTO can). Rotate the tuning shaft back and forth a few times—it should be easy to turn at this point, even with greasy fingers. Work the cam followers in and out about an eighth-inch (3-mm) or so and lubricate them. Grease the cam surfaces. You can now replace the PTO cover.*

reassembly

Replace and reconnect the PTO. Oil the turns counter, located between the PTO and the front panel. It should be possible, from the front panel, to turn the dry shaft with bare fingers. Grease the vernier knob assembly, gears and bearings, and remount the knob. Reset the knob to 14000 kHz, using the crystal calibrator to make sure the receiver is actually *tuned* to 14000 kHz, and try it out. Feels like a new receiver, doesn't it?

frequency jump

If your 75A4 was one of those that suffered from this annoying problem before, it should be gone now. The explanation is that the cam follower in a dried out, sticky PTO no longer rode easily on the cam. Instead, when the cam pitch changed slightly the follower hung up on dust off the cam, later dropping into proper position and causing that annoying jump in frequency.

One final caution. Keep your eyes open, both inside the PTO and around the drive train and dial mechanism, for dried grease, dirt, metal chips, galled surfaces, loose rivets or screws, or misaligned shafts or bearings. These can all be taken care of much more easily now, when the receiver is all apart, than they can late some night during the middle of the DX contest!

*A previous article on servicing Collins 51J series PTOs[2] has several worthwhile suggestions that apply to Collins 75A-series receivers as well. One of these is to replace the relatively unreliable tubular ceramic bypass capacitors in the PTO with disc ceramics, an easy job with the PTO removed from the receiver. editor

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1. Paul D. Rockwell, W3AFM, "Station Design for DX," *QST*, November, 1966, page 53.
2. William I. Orr, W6SAI, "New Life for the Collins 51J Receiver VFO," *ham radio* December, 1969, page 36.

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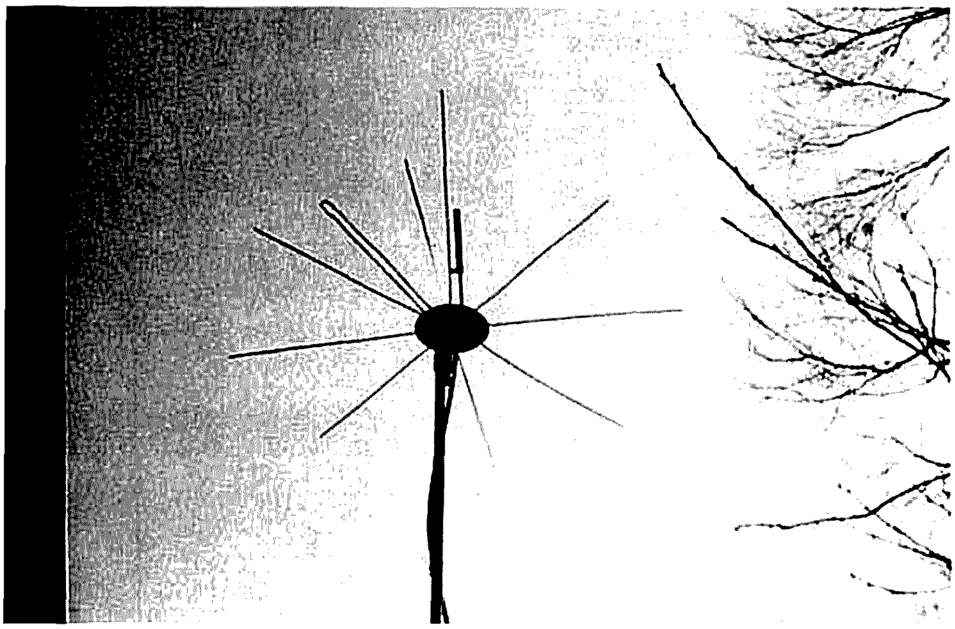

circularly-polarized ground-plane antenna for satellite communications

Combining the characteristics of two popular satellite antennas to yield a novel design optimized for satellite communications

Dale W. Covington, K4GSX, Marietta, Georgia 30060

Signal fading is a frequent source of frustration in amateur satellite communications. Much of the fading can be attributed to foibles in the patterns of simple ground-station antennas. The search for a stationary antenna with improved pattern characteristics led to the development of a circularly-polarized ground-plane antenna. This is a novel design that combines the best antifading features of two antennas often used in satellite work – the turnstile and the tilted-vertical, ground plane.[1]

The discussion of the circularly-polarized, ground-plane antenna has been organized in the following fashion: Two theoretical sections present the basic concept and the computed patterns. The next section examines certain problem areas which arise in achieving a practical antenna. Details of the construction, tune-up and operation of a two-meter prototype provide concrete illustrations of the design concepts.

the basic idea

Before plunging into the theoretical aspects of antenna design, let's list some general requirements placed upon fixed antennas used to communicate with OSCAR satellites. As indicated in the *appendix,* undesirable fading can be reduced if the ground-station antenna provides a good response overhead while focusing additional energy near the horizon where path losses are greater. Furthermore, vertical plane patterns should be independent of bearing azimuth, and the antenna should preferably be circularly polarized. The basic concept developed below is that each of these requirements is approached by a simple arrangement of two *tilted-vertical* antennas.

In discussing satellite applications for fixed antennas, it is convenient to map the far field radiation on an imaginary hemisphere centered over the antenna. Grid coordinates locating any observation point on the hemisphere are designated by an azimuth angle and an elevation angle. **Fig. 1A** shows a *far field hemisphere* over a quarter-wavelength vertical erected above a perfectly conducting ground plane. Vectors indicating the electric field magnitude and direction at selected points on the hemisphere are represented by arrows.

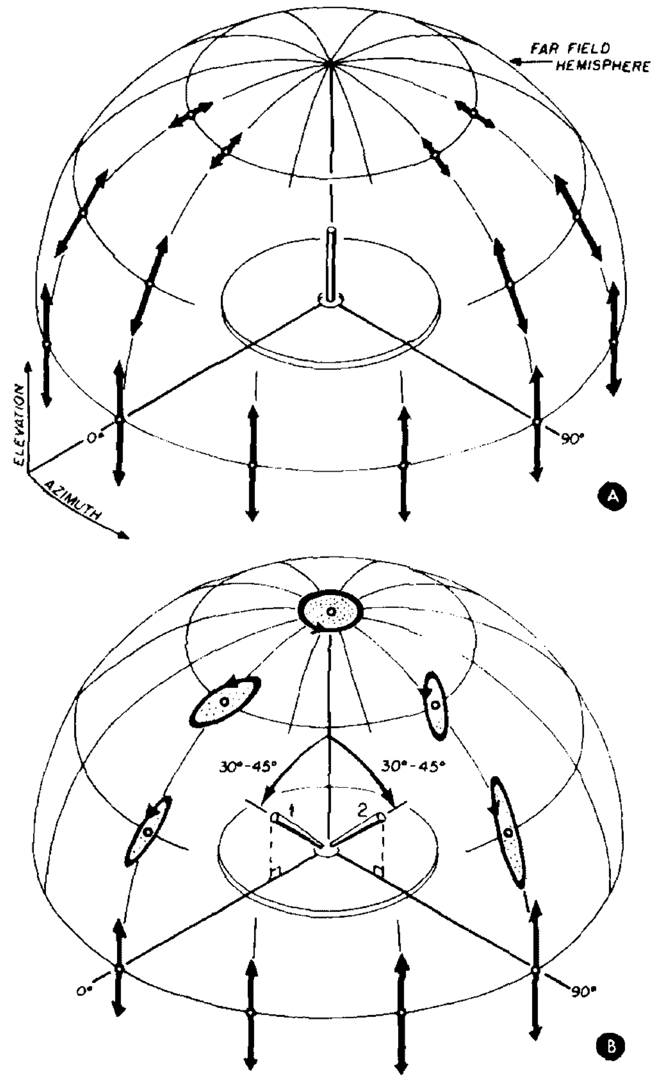

fig. 1. Isometric views of a single quarter-wavelength vertical (A) and two tilted verticals (B) above perfectly conducting ground planes. Grid coordinates marked on the far-field hemisphere locate directions in space above the antennas. Loci of the vector components of the radiated fields have been drawn at selected coordinate points.

While the magnitude of the electric field is independent of azimuthal bearing, it does vary with elevation angle. Notice how small the vectors become as elevation angles increase. The conventional plot of this effect is shown in the vertical-plane pattern in **fig. 2**. Here the dotted line plots the declining field intensity at elevation angles near 90 degrees. The pattern null directly overhead can be eliminated if the vertical is tilted away from the normal. *The heavy solid line in* **fig. 2** illustrates the relative pattern for a vertical antenna tilted at a 45-degree angle. In fact, this latter pattern does a good job of meeting the first two requirements listed above for ground-station antennas.

Returning to **fig. 1A**, observe that the electric-field vector is confined to oscillate in a plane containing the antenna and the observation point. The radiation under these conditions is linearly polar-

ized. A maximum transfer of energy will occur if the polarization of the ground-station antenna matches that of the satellite antenna. However, a deep fade will develop if a linearly polarized wave emitted by the satellite, for example, becomes oriented along a line of constant elevation angle instead of a line of constant azimuth angle. Such a turn of events could be introduced by satellite spin and/or Faraday rotation. The cure is to make either the ground or satellite antenna sensitive to fields oriented along any angle lying in a plane perpendicular to the propagation direction. A turnstile antenna achieves this characteristic if the excitation currents for the two perpendicularly crossed dipoles are of equal magnitude and in phase quadrature.

Fig. 1B gives an indication of the electric field components for two tilted verticals located in perpendicular planes. These quarter-wavelength verticals are fed by currents that are equal in magnitude but 90 degrees out of phase. For the special case where observation points lie on the horizon, the field is linearly polarized along the hemispheric meridians. At other elevation angles the tip of the instantaneous electric-field vector is generally not confined to oscillate in a meridian plane. Instead, the locus of its motion describes an ellipse. Directly over the antenna the ellipse degenerates into a circle. The important point is that the elliptical polarization of this antenna offers a degree of freedom from the undesirable effects of rotation of the plane of polarization while maintaining desirable, tilted-vertical behavior in the vertical plane radiation patterns.

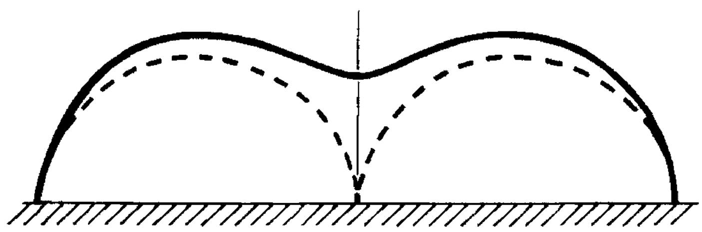

fig. 2. Relative vertical plane patterns for quarter-wavelength verticals. Dotted line: vertical antenna normal to perfectly conducting ground plane. Solid line: vertical antenna tilted away from normal by 45 degrees.

computed patterns

Mathematical expressions were derived which yield the vector components of the field radiated by thin tilted verticals erected above a perfectly conducting ground. Input variables for the expressions included the angular designation of the observation point, lengths and tilt angles of the radiators, and the amplitude and phase of the excitation currents.

The results of a series of numerical calculations using these expressions are presented below for the antenna shown in fig. 1B. Tilt angles of 45 degrees were chosen for the quarter-wavelength radiators. While current amplitudes were identical for the two radiators, the current flowing on radiator 2 was adjusted to lag the current flowing on radiator 1 by 90 degrees.

Fig. 3 gives the transverse projection of the locus of the electric field vector for discrete observation points on the far field hemisphere. The points are spaced around the hemisphere at increments of 45 degrees in azimuth and 30 degrees in elevation.

As **fig. 3** indicates, in general the field from the antenna is elliptically polarized. Right on the horizon, however, the field becomes linearly polarized. There is some dependence here upon azimuthal bearing. This is shown by the change in arrow length as the antenna is encircled and by the variation in the azimuthal pattern which is plotted in **fig. 4A**. The field strength on the horizon improves by 5.25 dB in moving from the position of minimum to maximum field. **Fig. 3** also reveals that the polarization sense is largely right-handed for outwardly propa-

gating waves although one quadrant of the hemisphere contains significant amounts of left-handed polarization. This reversal is denoted by the reversed rotation of the field vector loci.

Changing the 90-degree phase shift of the excitation current for radiator 2 from lagging to leading reflects the radiation pattern of fig. 3 through the vertical plane which bisects the angle between the

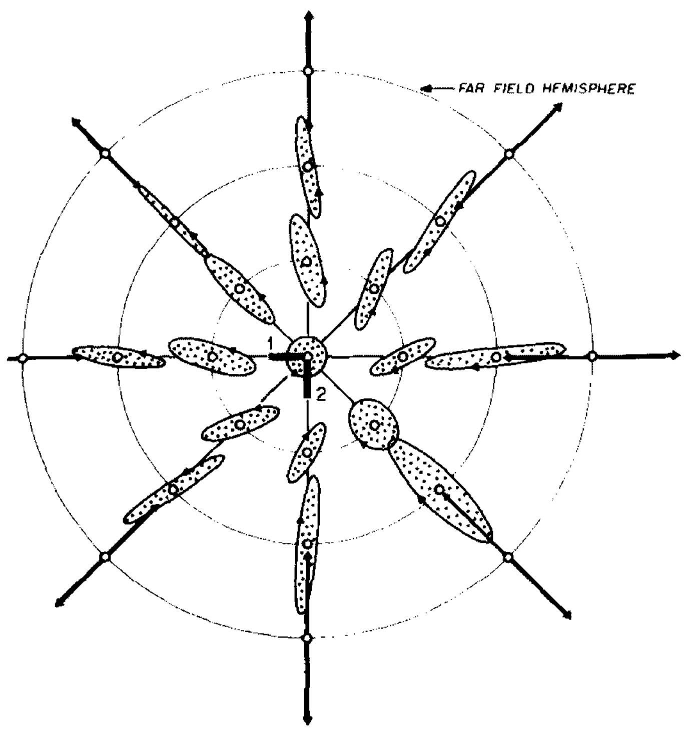

fig. 3. Far-field radiation pattern for the antenna shown in fig. 1B. Each radiator is tilted away from the normal by an angle of 45 degrees. The current exciting radiator 2 lags the current exciting radiator 1 by 90 degrees. Exciting currents are equal in magnitude.

two radiators. This means that the field polarization directly over the antenna changes from right-handed circular to left-handed circular. Fig. 4 gives an example of the azimuthal pattern reflection arising from a relative phase reversal in the excitation currents.

practical considerations

The theory discussed so far deals with ground-based, quarter-wavelength verticals. However, planting verticals for the two-meter band among the roses and tulips in the backyard flower bed doesn't hold particular promise in raising satellite contacts. The practical alternative at short wavelengths is to simulate the ground with a plane of quarter-wavelength radials. The resulting ground-plane antenna can then be installed in the clear where the electrical properties of local terrain features are less influential. It is difficult to evaluate the impact of such construction upon antenna radiation using simple theoretical models. Relatively little work has been reported which includes the effects of waves reflected from real ground beneath elevated ground-plane antennas cut for the satellite frequencies.

Experimental patterns of isolated verticals with limited ground planes exhibit the general characteristics of the ideal model where ground is infinite in extent and conductivity.[2] The principal deviation in practice occurs as slightly enhanced radiation at high elevation angles and slightly reduced radiation at the horizon. Therefore, a reasonable conjecture is that the fundamental framework of the radiation pattern shown in fig. 3 remains essentially intact after the ground radials are introduced. Some experimental results supporting this premise are presented in a subsequent section.

Interesting matching problems were posed by the constraints placed upon radiator currents I_1 and I_2. The 90-degree phase shift is conveniently obtained with a quarter-wavelength section of transmission line. Equal currents require careful selection of impedance levels at each end of the phasing line. Since the radiation resistance at resonance for a thin, quarter-wavelength vertical tilted by 30 to 45 degrees is of the order of 25 ohms, and the impedances of popular coaxial lines lie near 50 and 75 ohms, some impedance juggling has to be done.

Three possible approaches were considered. They are outlined schematically in fig. 5. At first glance the mechanical simplicity of fig. 5A is appealing. A phasing section of 26-ohm line is formed by paralleling two lengths of 52-ohm

coax. The net impedance at the antenna input is roughly 13 ohms. This is stepped up to 52 ohms by a 1:4 toroidal transformer.[3] A definite drawback to **fig. 5A** is the lack of electrical tuning for trimming-up the radiator currents. Of course, some tuning could be accomplished by pruning the element lengths and varying the tilt angles. On balance, the design seems to be more suitable for low-frequency operation where lumped-circuit tuning elements could be used.

Fig. 5C supplies considerable tuning flexibility at the expense of greater mechanical complexity. Each gamma section is adjusted for a 50-ohm match, and a simple coaxial transformer matches the 25-ohm impedance of the antenna to a 52-ohm feedline.

construction details

Since ground-plane antennas have long been popular with amateurs, the construction of a circularly-polarized, ground

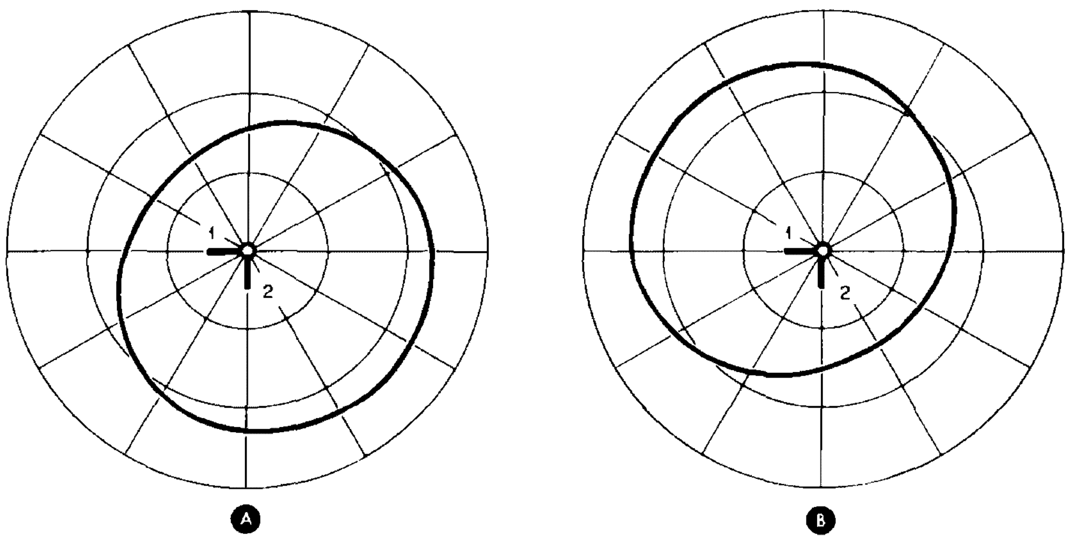

fig. 4. Azimuthal radiation patterns at elevation angles of zero degrees for the antenna drawn in fig. 1B. In (A) current exciting radiator 2 lags current exciting radiator 1 by 90 degrees. In (B) current exciting radiator 2 leads current exciting radiator 1 by 90 degrees.

The second design, **fig. 5B**, offers some tuning flexibility. When the gamma section is adjusted for 50 ohms, the 75-ohm, quarter-wavelength line transforms this value to an input impedance of 112.5 ohms:

$$Z_{in} = \frac{(Z_{line})^2}{Z_{load}} = \frac{(75)^2}{50} = 112.5 \text{ ohms}$$

If both conductors of the folded dipole radiator have the same diameter, then the dipole construction provides an impedance step-up by a factor of four:

$$Z_{dipole} = 4 \times 25 = 100 \text{ ohms}$$

Therefore, a good match to 52-ohm coax results by simply connecting the phasing line input and the dipole radiator in parallel.

plane poses no mystery once the basic design has been established. Formulas for the lengths of radiators, radials and quarter-wavelength coaxial lines at both hf and vhf have been listed recently[1] These formulas were used to determine the dimensions of a two meter prototype antenna based on the design given in **fig. 5B**.

Fig. 6 presents an exploded view of the gamma match[4] along with the dimensions of the two tilted radiators. The radiators are cut slightly longer than necessary from 1/8-inch (3-mm) diameter copper wire (number 8, B&S gauge). A threaded end (6-32 thread) of each radiator is fastened to a 6-inch (15.2-cm) diameter ground-plane disk with lock-washers and nuts. The remaining end of the dipole radiator and the center ele-

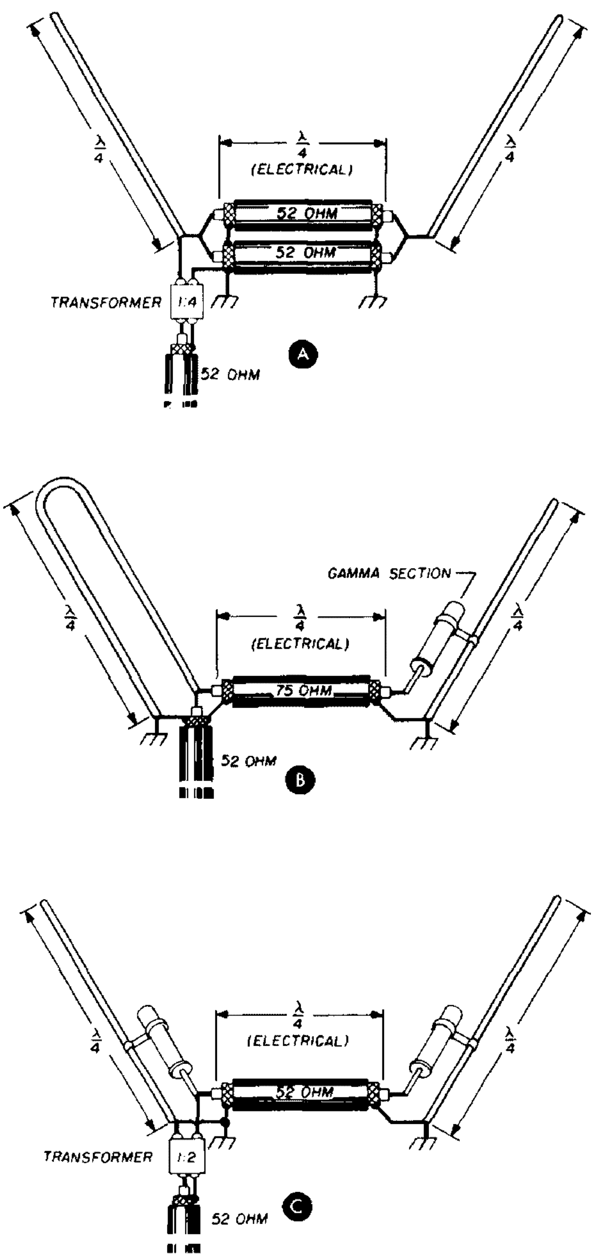

fig. 5. Three different ways to feed the tilted radiators of circularly-polarized, ground-plane antennas. Technique shown in (B) was adopted for the two-meter prototype antenna described in the text.

ment of the gamma match are connected to SO-239 coaxial panel connectors attached to the disk. These details are shown in fig. 7. The third SO-239 connector is used for attaching the feedline to the antenna. A tee would do this job better if the additional expense of the required connectors is no economic burden. Incidentally, the exposed upper portions of the SO-239 connectors should be protected from the weather with silicone sealant such as Dow Corning 3145.[5] Otherwise moisture will seep into the coaxial phasing line and the feedline.

Sixteen slits are cut around the outer edge of the disk to receive the eight ground-plane radials. The radials are also cut from 1/8-inch (3-mm) diameter copper wire. As **fig. 7** indicates, the disk is deformed slightly around each slit. This deformation not only holds the radials securely for soldering to the disk, it also adds considerable strength to the completed disk assembly. A right-angle bracket is a convenient way to attach the disk to a support mast.

tune-up

A simple tune-up procedure was devised for the two-meter antenna. The only instrumentation required is a transmitter and a vswr meter balanced for 50-ohm lines. Initially each radiator is individually pruned for resonance as indicated by a dip in vswr. Next, a 100-ohm carbon resistor is shunted across the dipole radiator, and the gamma section is installed on the single rod radiator. The vswr meter is connected to the gamma section input. The outer gamma tube and the sliding copper strap are alternately adjusted for a minimum vswr (below 1.3:1). The dipole resistor is removed, and the 75-ohm phasing line is then connected between the two radiators. A low vswr should now be observed at the antenna input (below 1.6:1). This figure may be improved by minor adjustments of the gamma section and the lengths of the phasing line and radiators. However, the primary reason for tuning adjustments at this stage is to balance the currents flowing in the radiators.

There are some simple ways to check the current balance. One check is to monitor the antenna vswr as the radiators are slightly detuned.[4] Under balanced conditions the vswr will increase by the same amount when a wire stub is clipped on to either radiator.

Another check is to measure the actual

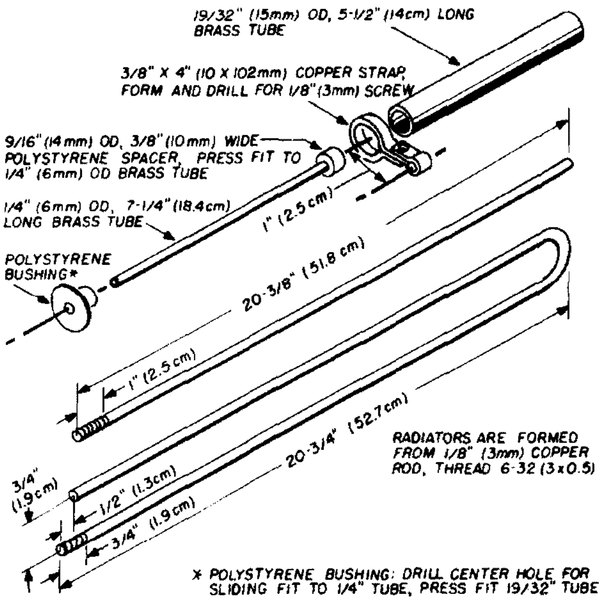

fig. 6. Dimensional details of the gamma match assembly and the two tilted radiators for the two-meter antenna.

radiation pattern of the antenna. Unambiguous pattern and polarization measurements are not easily made for low gain antennas. Yet there are two pattern characteristics that can be examined with an auxiliary dipole antenna coupled to a field-strength detector. Does rotation of the auxiliary dipole over the test antenna produce a highly elongated, dumbbell-type response? Does the vertically-polarized azimuthal pattern at low elevation angles show deep nulls which differ markedly from the anticipated patterns of fig. 4? If the test range is working properly, neither question will be answered affirmatively when balanced currents are flowing on the ground-plane antenna.

operation

Operating experiences with the antenna shown in fig. 7 appear to validate the pattern characteristics predicted by the theoretical model. Once the radiator currents were balanced, spot checks of the azimuthal pattern at low elevation angles revealed no deep nulls. Turning a dipole probe from the vertical to the horizontal position during these measurements indicated that the horizontally polarized components were down by at least 12 dB. A point of near circular polarization (variation of 1.4 dB as the probe dipole rotated through 360 degrees) was observed to lie within 15 degrees of the zenith.

A 20-watt transmitter connected to the antenna has been used successfully in establishing two-way contacts through OSCAR 6. It must be admitted in all candor that this is a marginal uplink arrangement if a high density of operators are concurrently using the satellite. AMSAT recommends radiated power levels of 80 to 100 watts for consistent satellite operation. The antenna gives particularly satisfying results during portions of orbits defined by large elevation angles. Near the zenith signals were strong and steady. As the satellite approaches the horizon, there is a gradual increase of signal fading characteristic of polarization rotation. Although signal levels fall at the lowest elevation angles where ground-plane limitations become important, the uplink signals are returned even when the satellite is less than 10 degrees above the horizon.

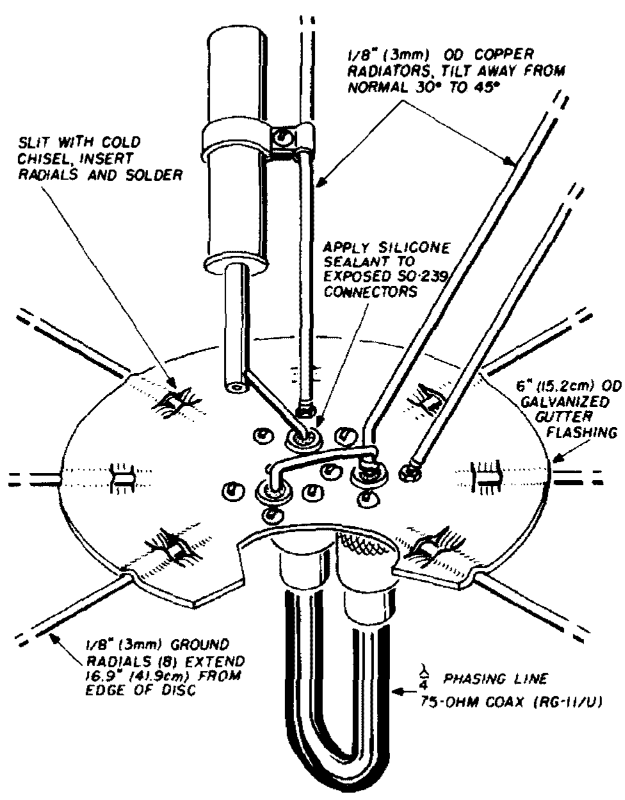

fig. 7. Assembly of a circularly-polarized, ground-plane antenna for the two-meter band. The electrical design is shown schematically in fig. 5B.

It stretches the imagination to come up with a single, fixed antenna which has all three of the pattern characteristics listed for ground-station antennas. The circularly-polarized, ground-plane provides a realistic compromise approach. The antenna does more right things than either turnstiles or tilted verticals. Moreover, building and implementing the ground-plane design is a very simple process — a lot simpler than trying to translate dreams of a tracking helix or crossed Yagi into fiscal and physical reality!

It is a pleasure to acknowledge that creative ideas and practical assistance were supplied by W4LKB during the construction and test phases.

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appendix

fading mechanisms

Many factors contribute to amplitude fluctuations in signal levels from OSCAR satellites revolving in circular, polar orbits. Some fading is inherent in the power-sharing feature of the linear satellite repeater. Other fading results from the changing geometric distance separating the ground observer and satellite. More complex fading is associated with the propagation of electromagnetic fields in an inhomogeneous and anisotropic ionosphere. The rich variety of the principle fading mechanisms is shown by the diverse entries in the first column of **table 1**:[6,7]

The second column in **table 1** gives some feeling for the physical conditions which enhance the individual fading mechanisms. Admittedly these statements are generalizations. Nevertheless, they are a useful guide in selecting antennas and operating conditions which minimize fading.

The last column in **table 1** lists several design techniques for improving the performance of ground station antennas. Examination of these techniques suggests three pattern characteristics that are desirable for stationary antennas: vertical plane pattern which is independent of azimuthal bearing, vertical plane pattern showing a gradual increase in radiation as elevation angles decrease and circular polarization.

A qualifying word should be added about the sense of circular polarization if both the satellite antenna and the ground station antenna are circularly polarized. Both antennas should be polarized in the same sense, i.e., right-handed, circular polarization (RHCP) or left-handed, circular polarization (LHCP). The vhf antennas planned for AMSAT OSCAR B will be circularly polarized. Once the satellite has stabilized, the correct polarization sense for ground station antennas are as shown below for stations in the Northern Hemisphere.

	mode	2-10 m repeater	432-145.9 MHz repeater	435.1 MHz beacon
polarization sense	transmitting	LHCP	RHCP	*
	receiving	*	RHCP	LHCP

The senses are reversed for stations in the Southern Hemisphere.

When a satellite using circularly-polarized antennas tumbles in space, the polarization sense becomes more difficult to define. In any event, the problem can be avoided for fixed ground-station antennas by installing a switch which selects the correct sense. The selection may be derived either from two separate antennas of opposite sense or from a single antenna which possesses sense reversal capabilities. The latter approach can be implemented

table 1. **Principal fading mechanisms encountered with OSCAR satellites.**

fading mechanism	conditions enhancing the fading mechanism	design techniques for ground station antennas which minimize fading
Operator loading	operators using excessive radiated power satellite moving in regions of the local sky visible to a high density of operators	position antenna for increased radiation along weak signal directions
Satellite moving over pattern nulls of fixed ground-station antennas	multi-lobed antenna patterns	select antenna type and height with a minimum number of pattern lobes switch in alternate antennas mechanically steer the antenna to track the satellite
Changing slant range	orbits passing over the local zenith	increase radiation at low elevation angles while maintaining some high-angle radiation
Spinning satellite	observer direction lies within a cone generated by a dipole antenna moving around the spin axis	use circular polarization use selectable antennas having orthogonal polarizations
	cross polarization between circularly-polarized ground and satellite antennas	switch polarization sense
Faraday rotation	low frequency daytime propagation along geomagnetic lines of force	use circular polarization use selectable antennas having orthogonal polarizations
Scintillation	low frequency nighttime ray paths traverse auroral zone ray paths traverse geomagnetic equatorial regions high sunspot activity	space diversity operation (this is not very practical for amateurs)
Ionospheric attenuation	low elevation angles low frequency daytime severe ionospheric disturbances	increase radiation at low elevation angles

rather easily by simply moving the transformer to the opposite radiator in the symmetrical designs (**fig. 5A** and **fig. 5C**) of the antenna discussed in the text.

The above comments place emphasis on fading control through proper antenna design. It is clear from **table 1** that careful choices of orbits, schedule times and operating frequencies also offer control over signal fading.

ham radio

three-digit

touch-tone decoder

for selective calling

Using tapped toroids to build a compact, low-cost Touch-Tone decoder

John F. Connors, W6AYZ, 3295 Brookdale Drive, Santa Clara, California 95051

With so many amateurs now occupying the limited number of channels in the vhf/uhf bands, the ability to be paged without having to continually listen to the chatter on the channel is a real asset. Since many amateurs can already transmit Touch-Tone signals, a reliable selective-call system can be built using the Touch-Tone approach. This article describes a Touch-Tone decoder that is suitable for a solo or group project and is inexpensive to build. It will allow the user to be alerted whenever his three-digit Touch-Tone number is received by his station by means of outputs which can ring bells, light lamps or enable speakers on receivers. The decoder is designed to operate on twelve volts, allowing mobile use, and it can be programmed to respond to any three-digit number.

basic design

Most commercial Touch-Tone decoders have separate filters for each tone channel to be decoded. The result is a large, expensive package. Since tapped toroid transformers are readily available,* I decided to use a single tapped coil for each of the two groups of tones recognized by the decoder. By pulling the

*Tapped Touch-Tone toroids can be salvaged from any scrapped (unrepairable) Touch-Tone pad, or purchased new from a number of commercial sources. Two such sources are: Aladdin Electronics, 701 Murfreesborough Road, Nashville, Tennessee 37210. L1 (low group) part number 426-847; L2 (high group) part number 426-848. Sangamo Electric Company, Communications Products Division, 11th and Converse Streets, Springfield, Illinois 62705. L1 (low group) part number 191983; L2 (high group) part number 191984.

appropriate tap to ground the coil can be tuned to the desired frequency.

In its initial state, the decoder toroids are tuned to the first digit pair of tones. Upon receipt of these tones the logic circuit switches taps on the toroid, tuning the decoder for the second digit pair of tones. Upon receipt of these tones the toroid taps are again switched and the decoder is tuned for the third digit tone pair. Upon receipt of the third tone pair the call latch is set and the decoder resets itself to the initial state; if only one or two of the three digits are received, the decoder will also reset.

The user may program the decoder for any three-digit number he wishes by rearranging the wiring between the two circuit boards. Six wires are used for this purpose. In this manner, each user on a channel or net can have his own private paging number within the group of 1000 possibilities.

circuit

The circuit was constructed on two boards. Board A contains the preamplifier, tone decoders and power supply. Board B contains the logic circuits to sequence digit recognition and provide the signalling output.

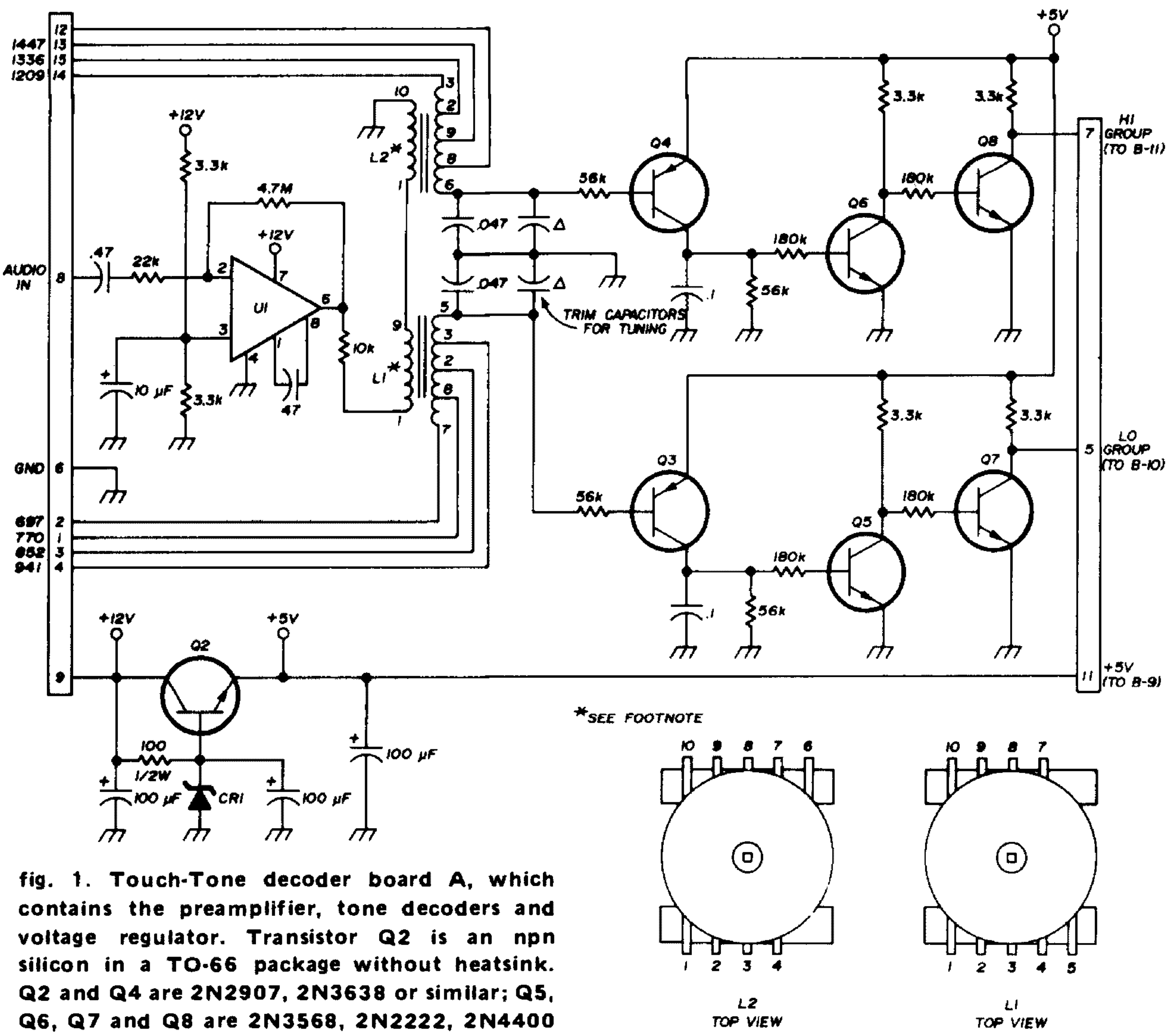

fig. 1. Touch-Tone decoder board A, which contains the preamplifier, tone decoders and voltage regulator. Transistor Q2 is an npn silicon in a TO-66 package without heatsink. Q2 and Q4 are 2N2907, 2N3638 or similar; Q5, Q6, Q7 and Q8 are 2N3568, 2N2222, 2N4400 or similar.

Audio is fed into board A (fig. 1) where it is amplified by U1 to drive the tone filters and decoders. L1 is the low group tone filter and L2 is the high group tone filter. The transistor drivers follow-

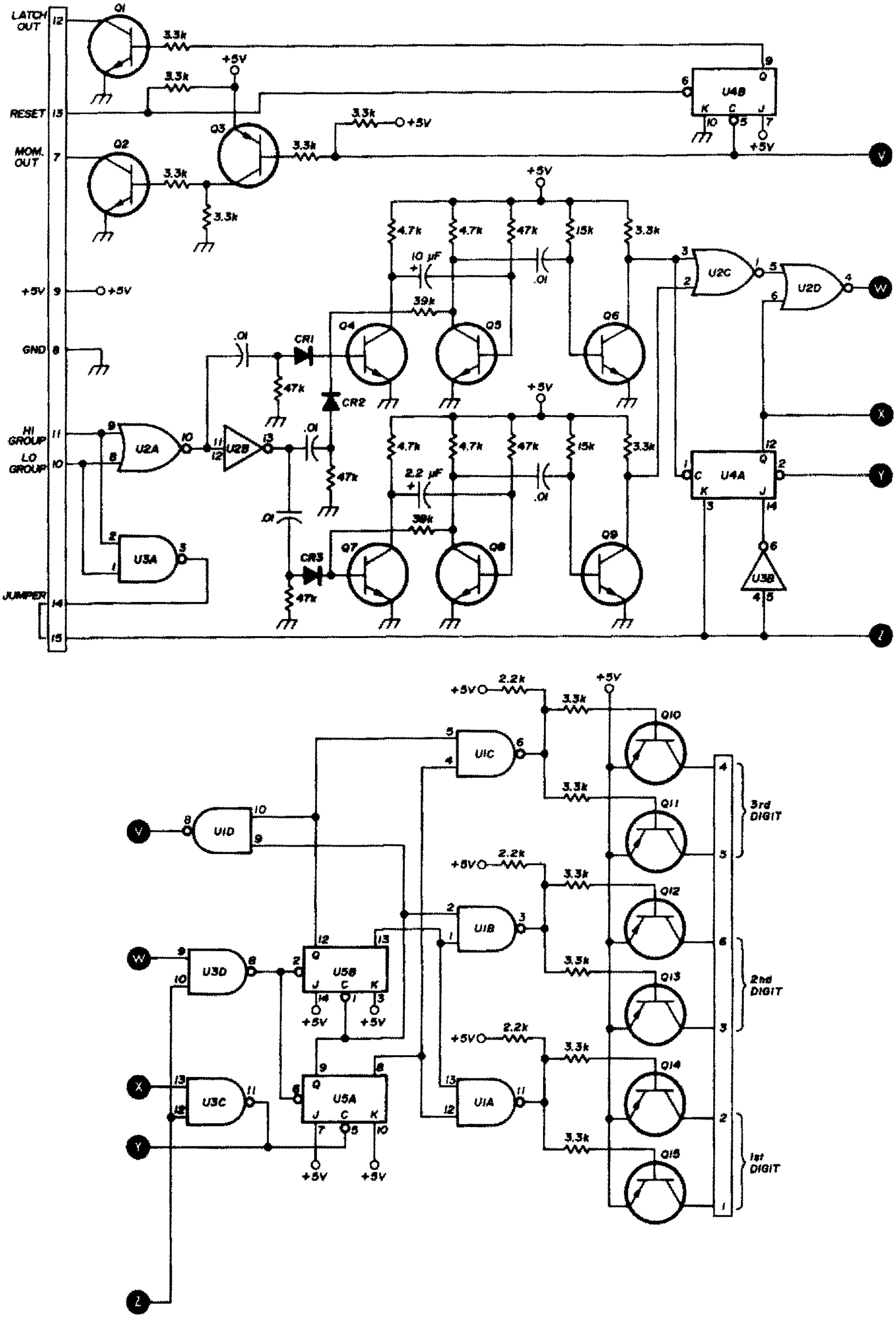

fig. 2. Touch-Tone decoder board B, which includes the logic circuits that sequence digit recognition, timing circuits and signal-activating output. Npn transistors are 2N3568, 2N2222, 2N4400 or similar; pnp devices are 2N2907, 2N3638, 2N4402 or similar.

ing the toroid filters convert the tones to standard logic levels. The power supply on board A (Q2) is a regulator to drop the 12-volt supply line down to 5 volts for the TTL logic.

The two logic signals, *hi group* and *low group*, are sent from board A to board B (fig. 2) where they are used to establish the sequencing of the decoder. The states

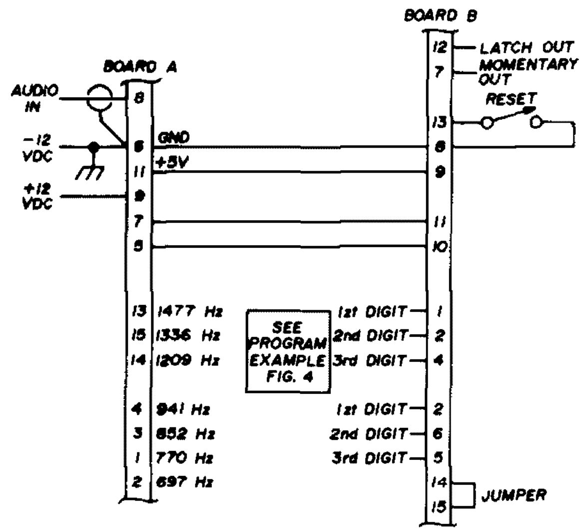

fig. 3. Interconnection wiring diagram between boards A and B. For programming example see fig. 4.

are sequenced by U5 which counts up from 00 as each digit is successively recognized. Gates U1A, U1B and U1C decode the output states of the counter and enable the transistor drivers (Q10-15) which pull the taps of the toroids to +5 volts. Transistors Q4-Q9 are time delays used to slow the circuit to a reasonable dialing rate and ensure the tones are legitimate before allowing the counter to sequence up. The upper time delay (Q4-Q6) is about 0.1 second and the lower time delay (Q7-Q9) is about 0.5 second in duration.

The output of gate U2A is true if either tone is present; the output of gate U3A is true only if both tones are present.

construction

The construction of the decoder is a matter of individual taste. I used two 3x3-inch (7.6x7.6-cm) printed-circuit cards which will plug into any standard edge connector. It is possible to use additional logic to reprogram the program wires connected between boards and change the selective call number remotely. This might prove handy for those who desire an extra command for some particular application.

To program a selective call number, six jumper wires are required. These wires are run between the open collector outputs of logic board B (fig. 2) and the open taps of the toroids on decoder board A (fig. 1). More than one collector will be connected to the same tap in cases where digits of the selective call number share the same row or column on the Touch-Tone keyboard. In effect, the collector outputs are logically being "ORed" by a parallel connection. This is permissible and will not affect the performance. A programming example is shown in fig. 4.

Fig. 3 shows the wiring for the connections between boards. The two outputs, *latch out* and *momentary out*, are open collectors which pull to ground in the true state. They are capable of sinking limited amounts of current (10 to 20 mA), so external drivers should be added if your particular application requires more current than that. The reset switch resets the entire decoder.

timing requirements

Each digit of the three-digit selective-

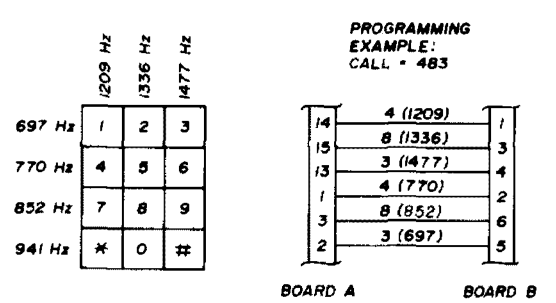

fig. 4. Touch-Tone tone matrix (left) and programming example. Note that common tones are used for different digits, and (for example) number 471 would have required that pins 1, 3 and 4 be strapped together on board B and connected to pin 14 of board A to provide response to a 1209-Hz tone for all three digits.

digit selective-call number must be transmitted for at least 0.5 seconds to be recognized by the decoder. Furthermore, there may not be a space between digits of more than 0.5 second or the decoder will reset. These requirements ensure good noise immunity and prevent triggering by voice or other signals on the channel.

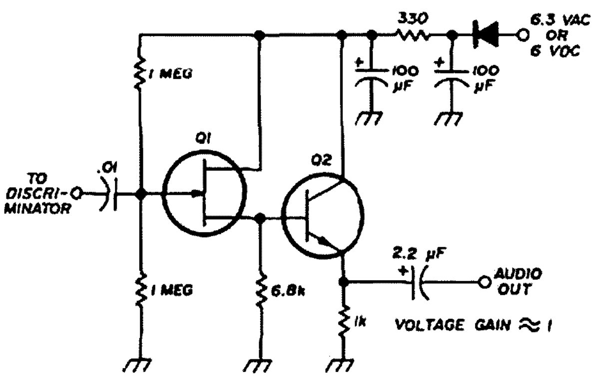

fig. 5. Circuit for simple discriminator amplifier, needed when receiver audio emphasis makes decoder response unreliable. Transistor Q2 is 2N3568, 2N2222, 2N4400 or similar.

other considerations

For best performance of a selective-call system the following factors must be considered. Poor frequency response in the audio sections of transmitters and receivers have been found to cause severe imbalance between the levels of the low and high tones in Touch-Tone systems. In some instances this difference could be great enough to create unreliable performance. If a stubborn case of *no workee* occurs, check out the audio response of the offending transmitter and receiver. A simple discriminator amplifier is shown in fig. 5 for those readers who would like to recover unprocessed audio from their receiver to operate the decoder without butchering the existing audio circuits.

For best results when transmitting Touch-Tone selective-call signals, acoustical coupling of the Touch-Tone audio into the live mike should be avoided, and the transmitter microphone should be disabled while transmitting the tones.

ham radio

how to convert your vtvm to an IC voltmeter

Simple solid-state vtvm conversion uses LM310H voltage-follower IC, costs less than five dollars

Mike Kaufman, K6VCI, 3733 Meadville Drive, Sherman Oaks, California 91403

The vacuum-tube voltmeter is probably the most common piece of test equipment used by amateurs. This article describes how to convert your vtvm into a battery-operated IC voltmeter (ICvm) at a total cost of about $4.00. Input impedance of the ICvm is identical to your original vtvm, and accuracy on the dc and resistance scales is identical. On the ac scales there may be a slight error at the lower ranges, although I have not verified this.

the circuit

Fig. 1 is a complete schematic of the ICvm. While designed around the popular Heathkit IM-11 vtvm, the circuitry is applicable to virtually all vacuum-tube voltmeters. Components in fig. 1 with three-digit call outs (i.e., CR100) are new, or are new applications for parts used in the original IM-11 vtvm (R133 for example, was R33 in the original circuit). The only new parts required are C100, CR100-CR103, S100, U100 and two 9-volt batteries (B100 and B101).

Begin the conversion by removing the pilot lamp, the ac line cord, the power supply transformer, capacitor and diode, the *ac balance* pot, the ac balance resistors, both vacuum tubes, tube bias components and all *zero adjust* components except the *zero adjust* pot itself. If you are modifying a Heathkit IM-11, the components to be removed are C1, R5, R10-R16, V1, V2, R24, C5, C6, R32-R35. Components R33, C5 and C6 will be used in the ICvm as will the circuit board and all the components remaining on it.

It is necessary to install a dpst toggle switch (S100) to operate as the new *on-off* switch. The vtvm switch wafer cannot be used as it has only a single pole. The new toggle switch (S100) may be installed on the front panel of the cabinet. Next, mount the two 9-volt transistor-radio batteries (B100 and B101). These batteries may be inserted into a battery holder, or they may simply be tied, with lacing cord, to the metal bracket holding the 1.5-volt battery. The 9-volt batteries may be connected to S100 at this time. Incidentally, rather than buy connectors for B100 and B101, make your own by removing the tops from two old 9-volt transistor-radio batteries and soldering a length of wire to each terminal.

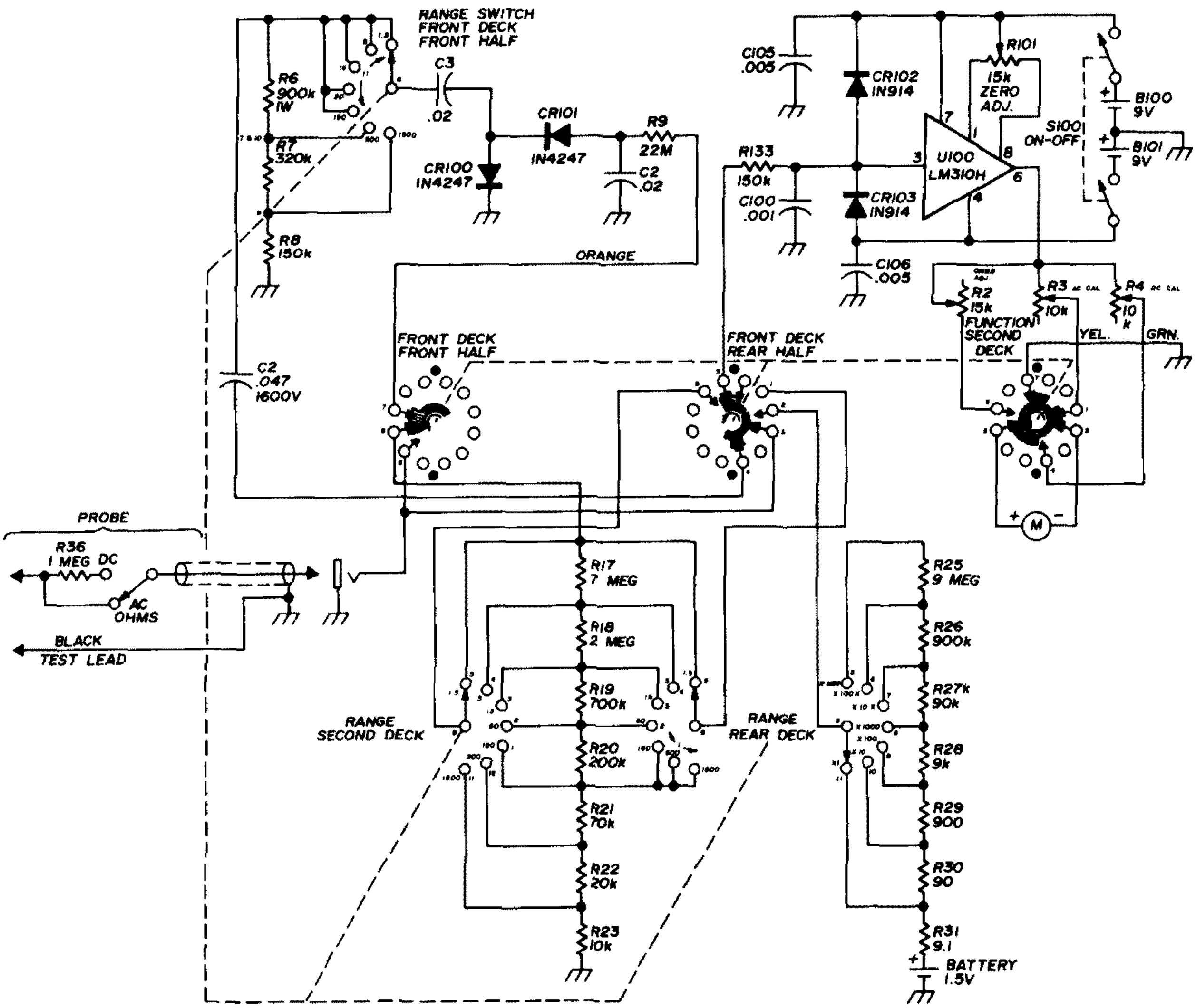

fig. 1. Basic IC voltmeter circuit may be easily adapted to any commercial vtvm although circuit shown here is the popular Heathkit IM-11. In this circuit two semiconductor diodes, CR100 and CR101, replace the original 6AL5 detector (see fig. 2). An LM310H high-impedance, unity-gain voltage follower, U100, replaces the original 12AU7 (see fig. 3).

Diodes CR100 and CR101 are solid-state replacements for the two diodes in the original vacuum-tube detector, V1, a 6AL5 (see fig. 2). Use the V1 tube socket to make the diode connections. Connect the cathode of CR100 to V1, pin 5, and the anode to V1, pin 2. Connect the cathode of CR101 to V1, pin 1 or pin 2, and the anode of CR101 to V1, pin 7.

Next wire in U100, the LM310H high input impedance (10^{10} ohms), unity-gain voltage follower.* This same IC may be used in virtually any vtvm (see fig. 3).

*The LM310H is available for $1.45 postpaid from International Electronics Limited, Post Office Box 1708, Monterey, California 93940.

The LM310H may be wired into the remaining tube socket, but remember to break all printed-circuit connections going to the socket. Wiring is not especially critical.

Capacitors C105 and C106 act to bypass the battery power supply and should be connected right at U100. Diodes CR102 and CR103 provide over-voltage protection in the event a large voltage is probed while the ICvm is switched to a low-voltage range. Regardless of how large a voltage is probed, CR102 and CR103 will limit the voltage at pin 3 of U100 to ±9 volts dc. Resistor R133 limits current into CR102 and CR103, and contributes to the protective circuitry.

Capacitor C100 is used to ensure that there is no ac at the input of U100. Although the value of C100 is not critical, increasing its value beyond .001 μF will introduce a noticeable time lag into your measurements. Resistor R101 is the original *zero adjust* pot.

Connect the *range* and *function* switches to the new circuitry, remembering to ground pin 7 of the second deck of the function switch. Your wiring should now be complete as shown in fig. 1. Install the batteries and the LM310H, and you are ready for calibration.

calibration

First ensure that the mechanical zero position of the meter pointer is correct. Then turn the ICvm *on* and adjust the *zero adjust* for either *dc-* or *dc+* zero reading with no probe input. There should be no appreciable change in the zero level when going from *dc−* to *dc+*. Next probe a known dc voltage and adjust the *dc cal* control to obtain the proper meter indication. Now put the function switch in the *ohms* position and

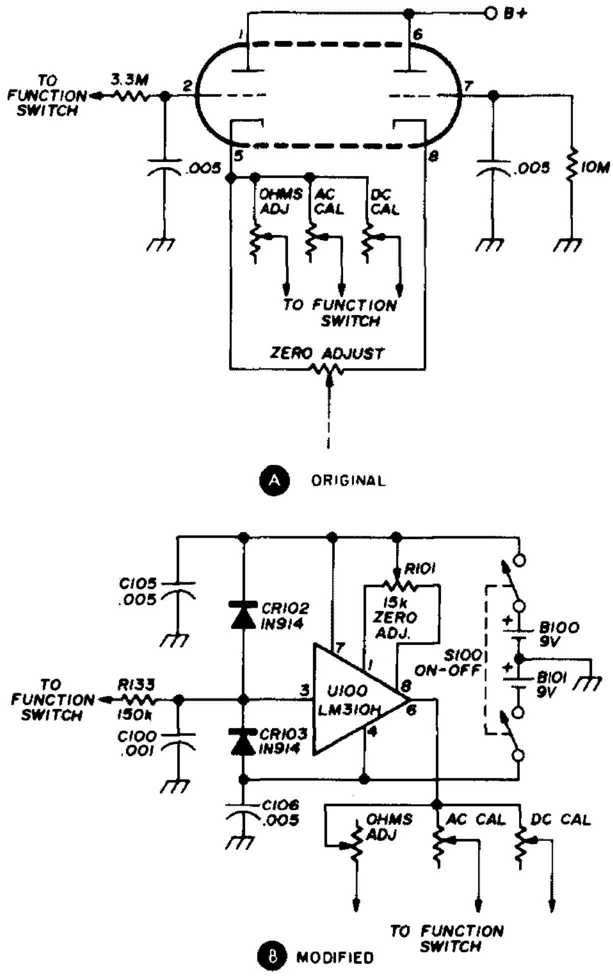

fig. 3. Solid-state replacement for the 12AU7 cathode follower uses a high-impedance, unity gain voltage-follower IC, the LM310H. The same circuit may be used to convert other dual-triode vtvm circuits.

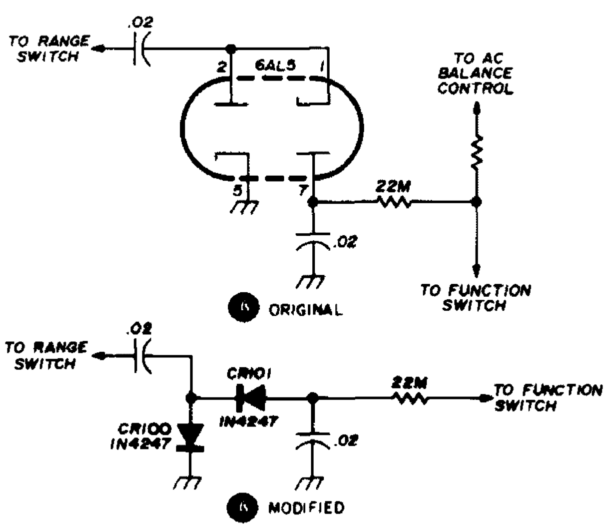

fig. 2. Conversion of the original vacuum-tube detector is simple and requires two diodes. Original ac balance control is not required in the solid-state version. Same circuit may be used with older instruments using 6H6 detectors as well.

set the *ohms adjust* control so that infinite resistance (probe open-circuited) registers full scale on the meter. Finally, put the function switch in the *ac* position and carefully adjust *ac cal* so that a known ac input (usually 117 Vac) registers properly on the meter. Unfortunately, it is rather difficult to obtain an accurate ac source voltage — nowadays the ac line is usually closer to 100 Vac than it is to 117 Vac.

The ICvm shown in fig. 1 has been in use for seven months with the original set of batteries. Since the LM310H draws about 4 mA from each 9-volt battery, it is well to remember to turn off the unit when it's not being used.

ham radio

vhfer's view of Solar Cycle 20

A discussion of Solar Cycle 20, now on the wane, and its effects on long-distance 50-MHz propagation

Pat Dyer, WA5IYX, 5315 Silvertip Drive, San Antonio, Texas 78228

As Solar Cycle 20 is finally, though somewhat unevenly, drawing to a close, it is appropriate to consider it and its effects on F_2 propagation in the spectrum above 30 MHz (and more particularly 6 meters). The following are the results of personal observations and data collection on both aspects.

As far back as the peak of Solar Cycle 19 in 1957 I had casually noticed sunspots, but it was not until mid 1963 that regular plots were made and records kept. The whole period since has involved daily plot-counts, weather and other factors permitting, using a 3-inch reflecting telescope (f/10) with a 60-power eyepiece to produce a 5-inch diameter projected image of the solar disk. Only when sky conditions were deemed suitable would a record be made, thus avoiding the inherent inaccuracies of trying to observe the sun through even moderate cirrus clouds, etc. The only large lack of data was the period from June through September, 1971.

Fig. 1 shows the daily average sunspot count by month. As some months may have involved as many as 25 or more plots and others as few as 10 or less, the smoothed averages shown in fig. **2** are more meaningful. These smoothed averages are made by taking sunspot data for six months before and after a given month and then averaging it. The so-called Wolf numbers from my data (made by taking ten times the number of sunspot groups and adding to the spot count) show substantially the same features as those presented in figs. **1** and **2**.

Unlike the official records made by the Swiss, this data shows a rather later

peak of Cycle 20 in 1970 vs the late 1968 or 1969 peak often cited elsewhere. A peak daily count occurred on November 17, 1970, with some 67 spots plotted in four groups. The Wolf-count peak for a given day was on February 1, 1968, with 53 spots in ten groups (W = 153). The results of this plot are shown in **fig. 3**.

The later stages of Solar Cycle 19 in 1963 are very evident in the graphs. This period was followed by a rather prolonged minimum running through 1964 and well into 1965. During these lean years the sun was spotless for many days at a time. A rebirth of activity was dramatic in 1966. In fact, the spotless sun of November, 3, 1966, was not duplicated until October 13, 1972. However, during the past year the incidences of zero counts have been becoming more and more frequent.

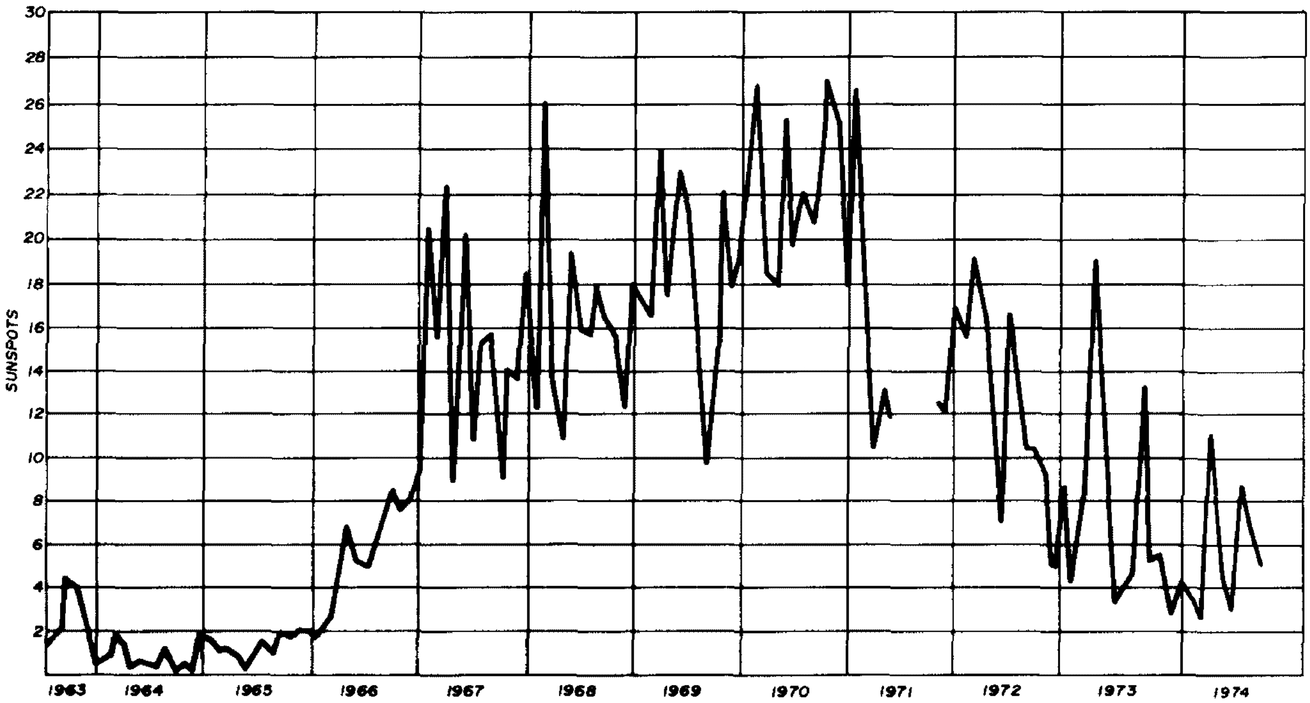

fig. 1. Raw daily average sunspot counts, 1963-1974.

Lacking more sophisticated equipment, I was unable to view the other associated solar events such as prominences and flares.

For those readers who are interested in conducting their own solar observations just about any astronomy book will provide the necessary information. My method of projection viewing and plotting is the simplest and the most safe. Photographic setups provide the most accurate record but the cost factor there can be limiting. Regardless of the method you use, *do not observe the sun directly without adequate filtering devices*. Both visible and infrared, as well as ultraviolet rays, must be reduced to safe levels to prevent permanent eye damage (which can occur quickly and painlessly).

vhf propagation

The ionospheric effects of a solar cycle depend greatly on the location of the observer. My interest in vhf propagation came about in the early 1960s first in the realm of TV-DXing and sporadic-E. I became an amateur in late 1963, and the results of 50-MHz E_s observations made during the period from 1964 to 1970 have appeared elsewhere.[1,2]

Much to my regret, suitable equipment for monitoring 30- through 50-MHz spectrum was not available until the fall of 1967. Even since then equipment has been on the simple side: a Radio Shack

Patrolman, and, in 1970, an Allied A-2586. Recently a Hallicrafters SX-62 has been revamped. Simple random-wire or whip antenna systems have been the rule.

On 50-MHz a low-power a-m transceiver and a five-element Yagi at 20 feet (6.1 meters) was used until the fall of 1968 when a higher power ssb rig was acquired.

Detailed records of F_2 MUFs in the 30- to 50-MHz region were not kept regularly until the fall of 1968. The late 1969, early 1970 data has been cut due to various receiver-related problems. Actually, the term MUF (maximum useable frequency) in these cases should be taken as MOF (maximum observed frequency) as no method (e.g., backscatter radar) was available to determine if the band was "open" higher than the highest incoming signal frequency.

Figs. **4** and 5 show the number of days each month that F_2 signals were observed in the contiguous United States and from Latin America at the indicated frequencies. As most of the latter are unidentified signals, there is a possibility that E_s propagation was inadvertently included at times. However, as will be discussed in more detail later, E_s often played a big role in providing link-ups with F_2 openings that would have otherwise passed, undetected. These graphs only include direct F_2 modes, and thus do not consider backscatter or transequatorial scatter (TE) propagation.

There was always a large difference in MUF behavior in the United States and Latin America, with little apparent relationship to one another. For instance, on many occasions South American signals were well above 45 MHz while in the United States even 10 meters was dead.

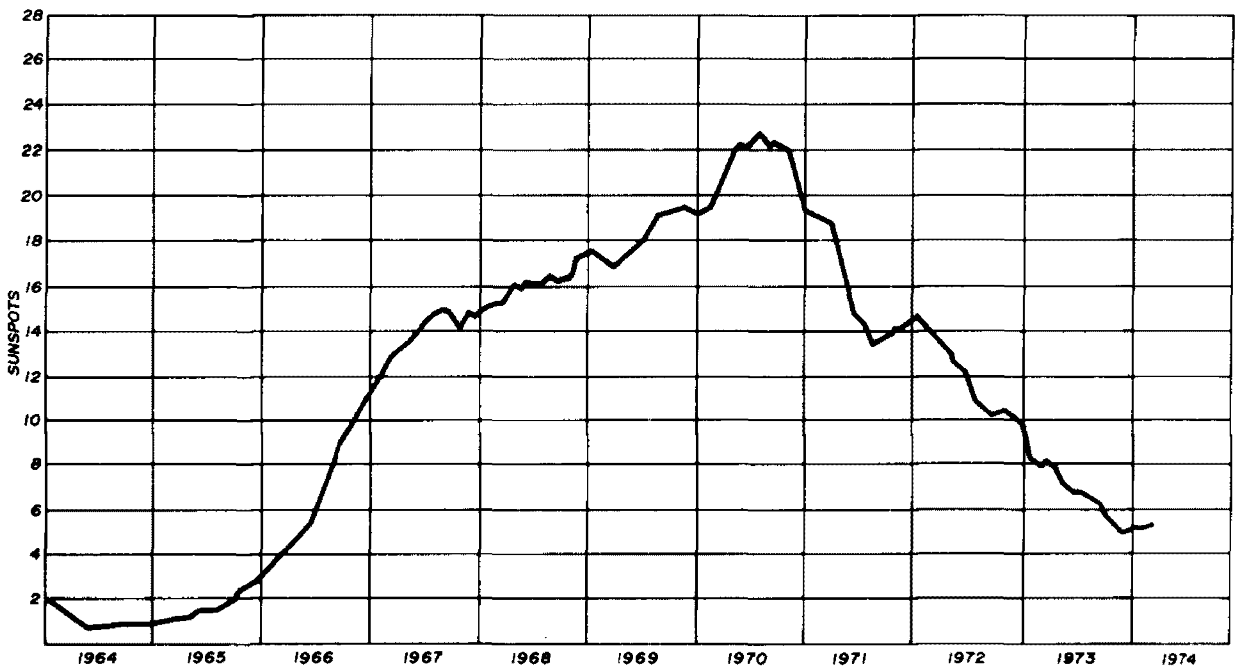

fig. 2. Smoothed sunspot counts, 1964-1973.

But, on other occasions the MUFs in the United States almost seemed to be keeping pace with those to Latin America.

In figs. **4** and 5 the F_2 "season" has been limited to September through April, although occasionally during the summer Latin American signals reached the 40-MHz region. Both seasonal and solar epoch variations are easily found. For U.S. MUFs the best months were October-December; this in striking contrast to the Latin American peaks of March and April. Year-to-year changes, while not always smooth, show the decline of Solar Cycle 20.

I should mention a word about the seemingly arbitrary frequency divisions used in **figs. 4** and 5. The selection is natural for the U.S. as FCC assignments

produce large groups at certain frequencies (i.e., fire departments at 33 MHz; pagers and mobile phones at 35 MHz; and law enforcement at 37 and 39 MHz). The Latin American situation is different as it is next to impossible to obtain station assignment information. Thus, no simple groupings are known which could make a more meaningful frequency division system than that used in figs. 4 and 5.

six meters

Since 50-MHz DX is of considerable interest to the vhf operator, it's worthwhile to take a more detailed look at Solar Cycle 20's F_2 effects on six meters. Table 1 gives a month-by-month summary of the number of days and minutes total open on 50 MHz by various modes. The mode determination is a rather simple process of considering the distances, peak antenna headings, fade rates, etc.

Fig. 6 shows the time of day of F_2 and TE openings on 50 MHz for the month of April summed over the period from 1967

table 1. Observed 50 MHz band openings.

		number of days (opening, total minutes)		
		F2	F2 bs	TE
1967	April	1 (85)	-	-
	September	1 (55)	-	-
	October	-	1 (15)	-
1968	March	3 (80)	-	-
	April	14 (635)	6 (530)	3 (40)
	May	2 (30)	-	1 (30)
	September	-	-	3 (255)
	October	-	1 (5)	1 (80)
1969	February	1 (40)	1 (30)	-
	March	1 (5)	3 (410)	-
	April	11 (290)	6 (655)	4 (60)
	September	-	-	1 (90)
1970	February	-	-	3 (150)
	March	3 (45)	1 (135)	-
	April	9 (340)	6 (265)	3 (200)
	May	2 (40)	-	-
	November	-	-	1 (10)
1971	March	1 (20)	2 (45)	-
1972	March	3 (60)	4 (170)	-
	April	8 (220)	3 (175)	-
	September	1 (10)	1 (45)	-
1973	April	1 (5)	1 (20)	-
	September	1 (15)	1 (20)	-
	October	1 (45)	-	-
1974	March	2 (35)	-	-
	September	1 (15)	-	-

to 1973. The time to be on the air is clearly in the afternoon. Almost without exception, I suspect that all the trans-equatorial scatter openings made it this far north with the help of an E_s link. The use of beacons by CE3QG and OA4C in those years was a priceless asset.[3] The lack of TE since 1970 is believed to be due, in large part, to the loss of activity from these two stations.

Backscatter, although not plotted in fig. 6, has much the same shape with earlier onset and later fadeout points. This is very consistent with the pattern of F2 backscatter from the southeast, followed by direct F2 from South America proper, ending with backscatter again from the South and Southwest.

The 50-MHz F2 paths to South America's more remote end, namely Argentina, Uruguay and Chile, are very likely the result of what are known as F2-F2 paths, shown in fig. 7. These are sometimes called trapezoidal paths due to their shape, and they provide very strong signals since an intermediate ground reflection with signal loss is eliminated. The geomagnetic equator, with its attendant "bulges" of F2 ionization on each side, is responsible for these tilted layers.

The geometry of the F2-F2 path is likely a rather ticklish affair requiring several different conditions to coincide. For example, if the ionization on the more northerly bulge of the path is not correct, the path is disrupted. Too low a level will cause the 50-MHz signal to overshoot the second bulge to the south, while the level which is too high may cause undershooting. This may explain the often observed oddity of six-meter stations from Argentina and Uruguay appearing when all the stations in Ecuador and Venezuela were at 44 to 46 MHz.

Sporadic-E, often seen as a friend in linking up with an F2 or TE opening, can just as easily ruin, by topside reflection, what would otherwise be a good path as shown in fig. 8. Since E_s may be partially transparent, the effect is very likely quite variable.

Six-meter F2 backscatter can be either single or double-hop in nature (perhaps

giving rise to a total path length of 9000 miles or more). The best earth reflection regions are over the oceans, or to the south and southwest of my station. This is ground backscatter and not direct backscatter from the ionized regions *per se*. E_s effects here are much the same as with the other two modes already discussed.

The following is an expansion with comments of the 50-MHz effects summarized in table 1. Suitable references are noted in the cases of major events.

1967 The month of April brought me my first meeting of 50-MHz F2 DX. It was more than five months before it was heard again.

1968 With some openings in March, April proved to be the best month of the Cycle, helped along by vast amounts of early season E_s, which also aided the first TE openings noted here. E_s also kept the F2 still alive well into May. The fall, though providing plenty of E_s-to-TE links, did not bring in the huge F2 openings that were anticipated by many operators.

1969 A very strong magnetic disturbance on February 2nd brought in backscatter here and several other modes elsewhere.[4] An E_s-to-F2 link later in the month provided some 40 minutes of the ZK1AA beacon. March and April, in contrast to the previous spring, brought in much more backscatter than direct F2. The lack of F2 was probably due to the poor E_s season.

1970 Good E_s in February permitted TE once again. Overall, April was better than expected, with many instances of the Cook Island beacon. However, the highlight of the year and the Cycle was on March 8th where, in one 90-minute period 50-MHz F2 backscatter was noted in some 16 states as far north as Illinois, with direct paths to Puerto Rico. This was the largest magnetic storm of Cycle 20 and occurred during a time of year when it would do the F2 layer the most

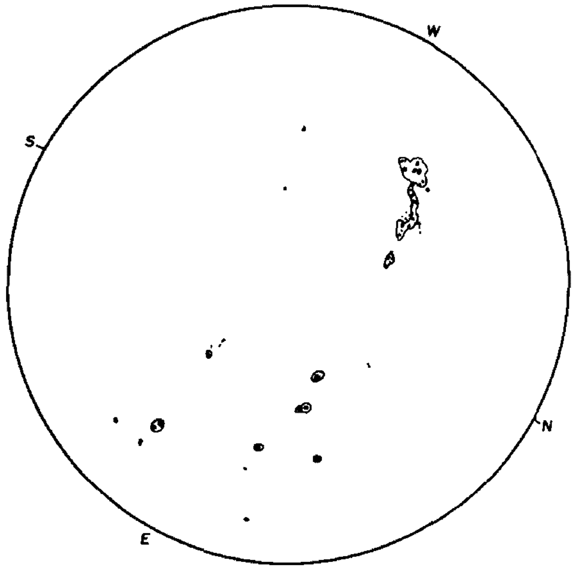

fig. 3. Sunspot diagram for February, 1968. Diagram shows 53 spots in 10 groups (Wolfe-number = 153).

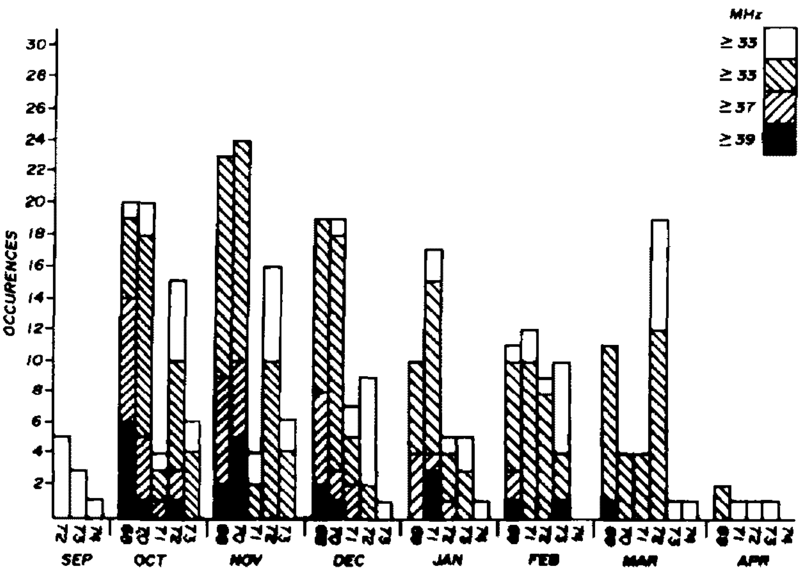

fig. 4. F2 MUFs for the United States plotted here were determined by noting the highest frequency incoming signal that originated within the 48 contiguous states. This meant the skip for a given frequency was down to 3000 km or less. When theoretical considerations are applied, the transcontinental MUF (e.g., W4-W6) was considerably higher than I could observe. Using the ITS maps, the 4000 Km MUF (nominal maximum 1-hop F2) can be extrapolated knowing how short the skip is on 35 MHz, etc.

good.[5] The last incidence of TE here occurred during November, in the midst of an E_s opening to Guatemala.

1971 50-MHz propagation might be best described as a recession, with only very scarce F2 effects in March.

northwest region, while second-hand reports from the 10-meter nets raved about the coast-to-coast 6-meter opening in progress.

Also, being this far south geomagnetically, no direct evidence of any of the many auroral events appeared. The closest incident was June 5, 1967, when a magnetic storm created extremely fluttery E_s (apparently) to Florida and perhaps either double-hop E_s or single-hop F2 to Puerto Rico. With so much E_s in June it is impossible to be sure of the modes without ionosonde evidence at hand. The prior week (May 25th) produced what were likely 49-MHz Latin American F2 signals while Florida had both visual and radio aurora.[7,8]

Along other lines, solar activity introduced vhf noise bursts to me in July, 1967. Although the event observed was nothing extraordinary, having the 50-MHz background noise rise by 40 or

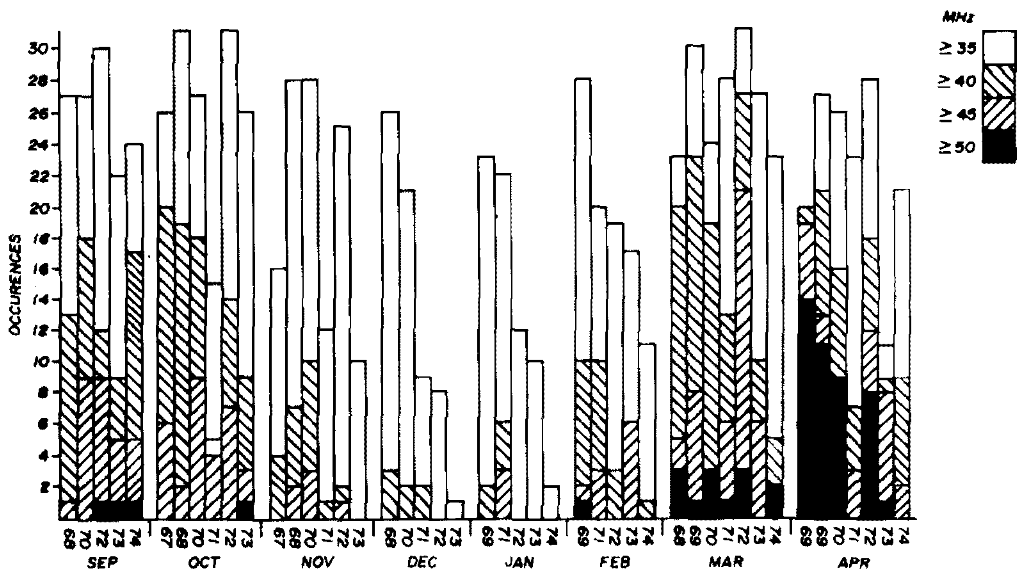

fig. 5. Latin American F2 MUFs.

1972 A rather unexpected upturn in solar activity in the spring provided the best spring March-April F2 in two years.

1973 Openings on 50 MHz, though very scarce, were amazing in that any occurred at all with such low solar levels.

Close followers of six-meter DX have probably noted by now a conspicuous absence of details on January 1, 1968.[6] While this location is very good for Latin American F2 (though not as good as Florida), when it comes to transcontinental F2 on 50 MHz it is just too close to each coast to get any. On the date in question 46-MHz was in from the Pacific

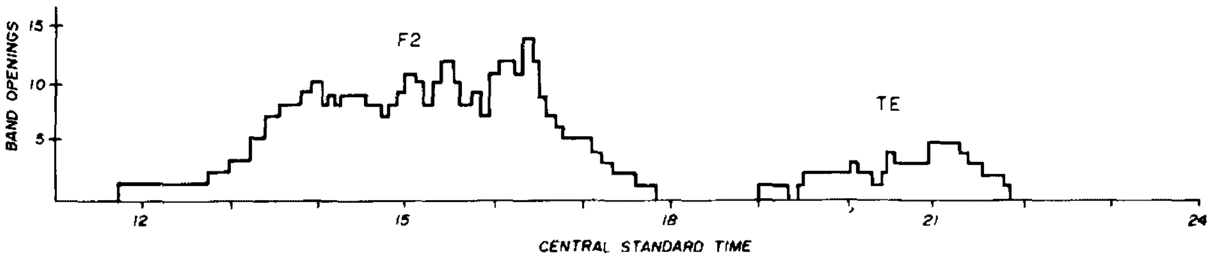

fig. 6. Graph showing 50-MHz F2 and TE openings during the month of April, 1967-1973.

50 dB for the first time was a memorable occurrence. During the ensuing years, while monitoring 30-50 MHz, numerous solar noise bursts have been logged incidental to MUF observations. A particular incident in April, 1973, when a solar noise burst was noted simultaneously with an increase in an F2 backscatter signal level, was vivid evidence of the association of flares, noise bursts and extra solar ionizing energy.

To step out of vhf for a moment, another equally dramatic trait of high solar levels is the high-frequency blackout (caused by extra D-layer ionization and consequent increase in collisions and absorption) when you are positive that your receiver has stopped working. Many of these blackouts were stumbled upon while attempting to get a WWV propagation forecast and the vhf E_s and F2 openings went along virtually unaffected.

future

While Solar Cycle 20 has not yet completely withered away, there is little doubt that it will be quite some time before the F2 effects on 50 MHz become as common as they were in 1968. However, devoted 50-MHz DXers might still be able to catch a few of the freak openings still left in the Cycle. For a better chance at catching the openings, the following suggestions are offered:

1. If you don't already have a receiver that will tune 30 to 50 MHz, by all means get one that does. While an SP-600 or one of its relatives is best, you can get by with a lot less.

2. Become familiar with the DX signals that frequent your area on the band. This can be helpful in looking for the more common E_s openings that might affect 6 meters.[9] When the conditions appear favorable, don't just listen, call CQ. You may end up with a hoarse voice and not get a reply, but at least you tried. For those fortunate enough to have beacons, it will be a lot easier.

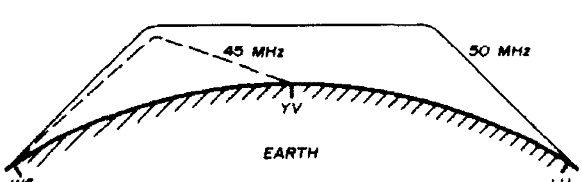

fig. 7. Propagation over an F2-F2 path from Texas to Argentina (not to scale).

3. Obtain copies of the *Telecommunications Research and Engineering Report 13*. This has ionospheric predictions for F2 median MUF (both at zero and 4000-km ranges) as well as normal E at various solar cycle levels (with interpolation methods) in the form of world maps at two-hour intervals for each month of the year.* While not all that useful in predicting the 50-MHz openings, they have supplanted the monthly bulletins that ITS (formerly CRPL) used to publish and are valuable in determining in which directions from your location the MUF is peaking at a given time of day, season, etc. The F2-F2 path can be inferred from the maps, but its MUF will be somewhat greater than the 4000-km MUF given on the maps for each "bulge" near the equator.

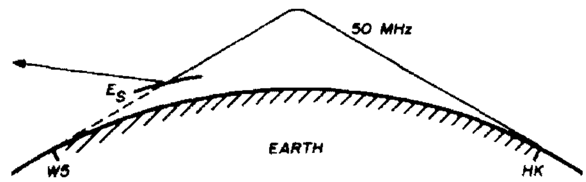

fig. 8. Sporadic-E shielding of a 50-MHz F2 path (not to scale).

4. If you have high-frequency capabilities, do all you can with contacts in regions where 50 MHz might be likely to stir their interest in at least listening on that band, if not actually setting up a station. Innumerable openings have been lost due to lack of 50-MHz activity in Venezuela and other parts of northern South America – openings where all sorts of high-frequency harmonics were pouring

*Four volumes are available. Volume 1 is the instruction manual ($.30), while Volumes 2, 3 and 4 are for smoothed sunspot levels of 10, 110 and 160, respectively ($3.00 each). Order from the Superintendent, U.S. Government Printing Office, Washington, DC 20402.

through on or near 50 MHz. In addition, the use of beacons on 50 MHz should be encouraged in these DX spots.

The foregoing and suggestions elsewhere[10] are a valid formula for getting into shape for the next solar cycle peak — you have plenty of time as it will not likely occur until the latter part of this decade. I'm afraid that a lot of plans for Solar Cycle 20 got going too late to be of much benefit, particularly the set-up of some 50-MHz beacons.

conclusion

I hope that this article will serve as a stimulus to others to undertake similar observations and recording of their data. This is only one of the ways amateurs can justify the portions of the spectrum we occupy — by contributing to the basic understanding of vhf propagation.

Over the years I have been indebted to several fellow amateurs for their encouragement and advice. Bob Cooper, W5KHT, deserves special acknowledgement for getting me to keep more accurate notes on the F2 DX conditions in the 30-50-MHz region. I wish that I had started earlier in the Cycle.

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ham radio

improving the performance of low-frequency vertical antennas

Three basic ideas for improving the efficiency of your antenna system

Raymond H. Griese, K6FD, 405 Giannini Drive, Santa Clara, California 95051

This article discusses certain ideas and methods that improve the efficiency of vertical antennas. If these methods are followed the resulting antenna will perform as it should, approaching textbook levels of efficiency.

The first idea to consider is that radiation from the antenna is a *natural* phenomenon. It is created by the changing current, either ac or pulsating dc, that is flowing in an electronic circuit. It's that simple. As a matter of interest, all electronic circuits radiate to some degree but the amount of radiated energy is so small it is ignored. An antenna is a special type of electronic circuit that maximizes radiation.

The second idea is the counterpart of the first: radiation can be suppressed only by mirror-image currents. This is the principle that is used in 300-ohm TV ribbon line to prevent radiation. Each conductor carries a current equal to the other in amplitude and frequency but 180° out of phase. The radiation from each conductor cancels that from the other, or very nearly so, thus for practical uses the line is considered non-radiating.

Considering the first and second ideas leads to the conclusion that antennas should not be built like a two-wire transmission line; they should be arranged so that there is no suppression of radiation by mirror-image currents. Fortunately there are two basic methods available to accomplish this goal. One is to spread the two conductors apart so that the radiated field from one conductor will not completely cancel that of the other. Examples of this type of antenna are loops, quads and some types of rhombics.

The other method is to use a single conductor in which current is maximized by tuning it to resonance at the operating frequency. This is possible because the current in the conductor is reflected from

the open end. The reflected current is in phase with the original current and its radiation adds to that of the original current. The familiar wire antennas – longwires, Zepps, Windoms and so on – use this technique and differ only in method of feed. It will be assumed that you are familiar with loops and single wires so they will not be discussed further. The grounded vertical version of the wire antenna will be discussed, however, in terms of its equivalent electrical circuit.

The antenna, shown in the form of its equivalent electrical circuit in **fig. 1** consists of a power source, a two-wire line and a resistive load which represents the radiation and loss resistances of the antenna. **Fig. 2** shows another form of the antenna circuit. Here, however, it must be remembered that the antenna is self-resonant at the operating frequency and presents a 73-ohm load (or thereabouts) to the line. Although diagrams such as these are helpful in understanding antenna operation, the circuit of **fig. 3** is even more helpful.

In fig. **3** the antenna is considered to be two quarter-wavelength antennas in series, the actual case. The connection between the ends of the antenna is fictitious, but this is actually the type of load the power source sees. At dc it will be an open circuit, but at radio frequencies it acts just exactly as if it were a

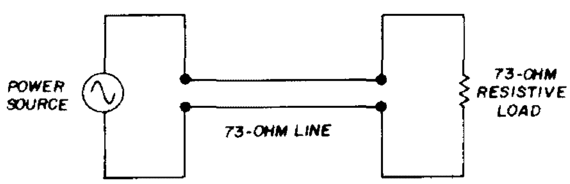


fig. 1. Equivalent electrical circuit of the antenna.

closed circuit. The standing wave of current in each quarter-wavelength antenna is responsible for this effect.

This circuit satisfies the dual requirements for the line and the antenna. The current *into* each quarter-wavelength antenna is 180° out of phase with the other, but one antenna is physically reversed 180°. This makes the two antenna currents in phase, maximizing radiation. *Line* currents, however, are 180° out of phase, minimizing radiation. This antenna is usually referred to as a half-wavelength, center-fed doublet or dipole. It is really two quarter-wavelength antennas operated in push-pull.

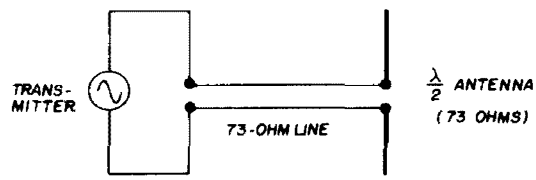


fig. 2. Common form of the antenna circuit.

Suppose the two quarter-wavelength antennas in push-pull are exchanged for one quarter-wavelength antenna and a ground connection. This circuit is usually drawn as shown in fig. **4**. Redrawn in the ac closed-circuit form **(fig. 5)** it looks exactly like **fig. 3** except for the value of the ground resistance. The ground resistance will be from about 2 to 200 ohms, depending on the physical arrangement of the ground system. A 2-ohm ground resistance is typical of a broadcast station ground system composed of 120 radials; a 200-ohm ground is a typical value for a ground rod in sandy soil.

Applying series circuit power calculations to the antenna-ground circuit brings out some very useful and interesting information. Power delivered to the antenna-ground circuit divides proportionally according to the value of the resistances. With a 36.5-ohm quarter-wavelength antenna, a 2-ohm ground absorbs 5% of the power, a 36.5 resistance absorbs 50% and a 200-ohm resistance will absorb 85% of the power. This indicates that there are only two ways to improve efficiency. The first is to reduce the ground resistance to as low a value as possible. The second is to raise the antenna resistance to its highest possible value. Surprising as it may seem, this *simple solution is a true engineering* solution to the problem of achieving efficiency in grounded vertical antennas.

Reducing the ground resistance by using multiple ground rods is poor practice. Current distribution beneath the quarter-wavelength vertical is such that ground rods do not intercept much of it. Ground wires are much better, either on top of the ground or immediately below the surface, and should be roughly a quarter-wavelength long. One radial is roughly 40 ohms; two radials, 180° apart,

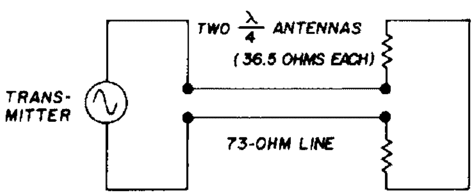


fig. 3. Ac equivalent of the antenna circuit.

are about 20 ohms; four radials get down to about 15 ohms. This is about the practical limit for amateur antennas – more radials are usually not worth the effort. It would, for example, take 116 more to get down to 2 ohms. Four radials should allow the antenna to operate at 70% efficiency – just 1.5 dB below maximum.

The radials don't have to be on or in the ground, they can be elevated above ground as well. This leads to the type of antenna known as the ground plane. One quarter-wavelength radial (antenna) exhibits a resistance of 36 ohms. If there was no interaction, two would be 18 ohms and four would be 9 ohms. However, the currents are 180° out of phase and the radiation is low, so the resistance is lower than you would suspect. I have not been able to determine the effective resistance of the ground plane itself, but suspect it is lower than 9 ohms.

Excellent results can be obtained with two or more radials on the ground-plane antenna. However, in emergencies only one radial will work. The advantage of the ground plane is that it can be elevated above ground, out of the vicinity of all neighborhood "hardware."

antenna resistance

Increasing the antenna resistance with respect to the ground system is also a good technique for improving antenna efficiency. For example, the feedpoint resistance of half-wavelength verticals runs from 500 to 3000 ohms – 500 ohms for towers that are wide compared to height and 3000 ohms for a very thin wire such as you might use for a baloon-supported antenna. A TV pipe mast, when used for an antenna, exhibits about 1000 ohms resistance. Considering this antenna on the basis of the third idea, it can be seen that any type of ground system will work well, including that 200-ohm ground rod! The major difficulty with the half-wavelength antenna is that it is twice as high as the quarter-wavelength antenna and the 1000 ohms or so input resistance is harder to match to a 50-ohm transmission line. This is especially true if you are running high power.

Remember that it isn't mandatory to operate at the quarter- or half-wavelength points. The advantage is that these antennas are self-resonant and easier to match. Antennas that are not self-resonant can be resonated by the addition of tuning coils and capacitors which usually makes matching more difficult. Short antennas can be loaded with coils part way up the antenna, the tops can be folded over, and so on. All of these techniques are designed to raise the radiation resistance of the antenna so it will accept a higher percentage of the power delivered to the antenna-ground circuit.

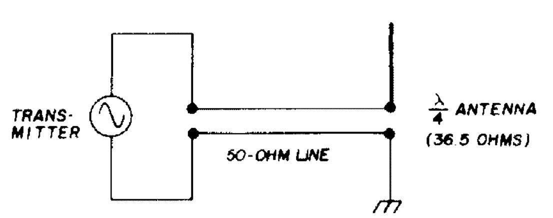


fig. 4. Common schematic for a vertical antenna.

Radials, too, can be loaded or tuned if space does not allow the use of radials a quarter-wavelength long. Radials can also be folded or bent quite severely without materially decreasing their effectiveness. Loading and folding are usually used on 1.8 and 3.5 MHz where full-size verticals

become physically large and difficult to erect on the normal city lot or apartment house roof. Many such arrangments are described in the textbooks that cover low-frequency radio engineering, and many of the old books from the spark era have a wealth of ideas for operating antennas on frequencies very much lower than the quarter-wavelength resonant frequency.

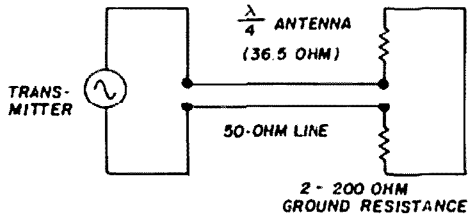


fig. 5. Ac equivalent of the vertical antenna circuit.

summary

To sum up the discussion for improving medium- and high-frequency vertical antennas, the following ideas should be thoroughly understood and put into practice:

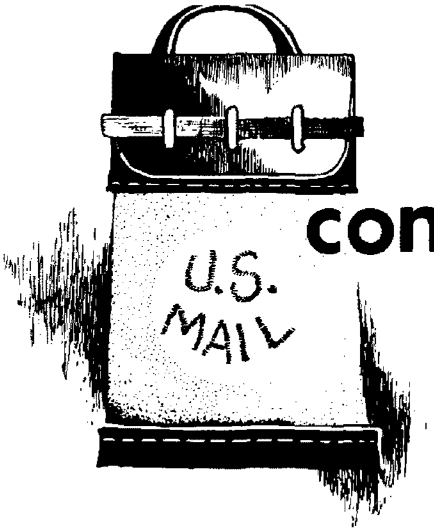
1. Ac current flowing in a conductor radiates energy. This is a natural attribute and requires no special expertise.

2. Radiation can be suppressed only by mirror-image currents flowing in nearby conductors or structures. Hence, the antenna should be erected away from or above these obstructions if possible.

3. The antenna and ground resistances should be arranged to maximize antenna resistance and minimize ground resistance.

These ideas and rules are not new. They are sound engineering principles that have been in existence since radio first came into use. However, they seem to have been neglected in most recent antenna articles. There is an infinite variety of ways vertical antennas can be built, and if the construction meets the requirements embodied in these three basic rules, you can be assured the antenna will work properly.

ham radio



comments

pi-network design aid

Dear HR:

My article on pi networks in the May, 1974, issue of *ham radio* (page 62) has caused some confusion because of an honest (but neglectful) error in the example using **eq. 1** for tube plate-load resistance. When the values given are plugged into the equation, the answer is 5000 ohms, not the 1800 ohms indicated. Since the table for the B&W 850A coil is for R_L = 1800 ohms, the values should have been E_B = 1900 volts and I_B = 0.525 A, which works out to 998 watts input for a plate-load resistance of 1810 ohms. These values would be typical for, say, a pair of 813 or similar tubes. However, the newer tubes designed for linear amplifier service use plate voltages on the order of 3000 to 4000 volts.

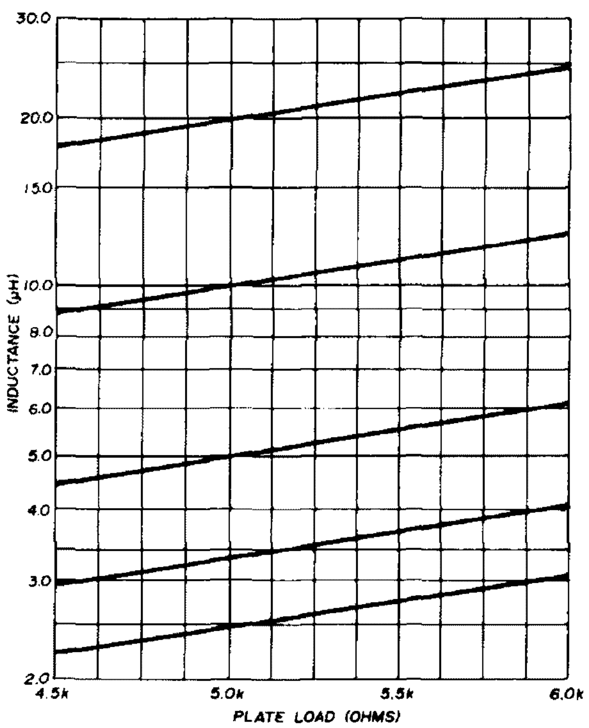


fig. 1. Curves of inductance vs plate load impedance for pi networks extrapolated to include values for high-power tubes.

The principles in the article are still valid, however. If the curves of inductance vs plate-load impedance are extrapolated to include the higher values (see **fig. 1**) it is clear that something must be done to the popular B&W 850A coil to obtain optimum inductance (assuming a nominal Q of 12).

Inductance for plate-load resistance, R_L = 5000 ohms

band	L(μH)	B&W 850A (μH)
3.5	20.0	13.50
7.0	10.0	6.50
14	5.0	1.75
21	3.3	1.00
28	2.5	0.80

The inductance values shown for the B&W 850A coil are those published in their data sheet. I have no axe to grind with B&W or their product, because the B&W 850A coil is well made and a pretty good bargain, even at today's prices. However, as I stated in the article, it is a design compromise, and to obtain optimum performance the coil must be modified to obtain the inductances shown.

Alf Wilson, W6NIF
Encinitas, California

memory keyer

Dear HR:

The electronic keyer with random-access memory described in the October, 1973, issue has generated considerable interest in this area. However, some operators have come to the conclusion that, "The (expletive deleted) thing doesn't work right." The first dot of a character following a character ending in a dash came out as a dash. Since the

keyer circuit is patterned essentially after the Micro-TO keyer, the problem isn't too difficult to solve — the toggle of the second J-K flip-flop should be connected to the Q output (pin 9) of the 7473, *not* the $\bar{Q}$ output (pin 8) as shown in the schematic. (The Q output also goes to the 7402 NOR gate B.) With this simple change the keyer works perfectly.

The modification is easily made if the PC board is used. Simply bend pin 8 of the 7473 away from the IC so it will not short to the socket, and short pins 8 and 9 on the PC board.

Howard M. Berlin, K3NEZ
Edgewood Arsenal, Maryland

kilowatt linear modifications

Dear HR:

Regarding my five-band kilowatt linear amplifier which appeared in the January, 1974, issue of *ham radio*, following are two modifications which will improve its operation. The first, suggested by W1OR, provides monitoring of both plate *and* screen currents. By removing the ground from the plate meter and the positive screen supply, and connecting the plate- and screen-current meters as shown in fig. 2, both can be monitored.

The second modification provides better plate and screen efficiency when using the amplifier on CW. By adding a second zener diode in series with the one shown in the original circuit, the static plate current may be reduced to near zero for operation on CW. When this second zener diode is shorted out, the idling plate current returns to the proper level for linear ssb operation. Note, however, that the switch is at −300 volts with respect to ground so it must be insulated from its frame by at least that amount.

Some readers are apparently having difficulty in determining the proper connections for the zener diode. When using a volt-ohmmeter on the ohms scale, the polarity is sometimes reversed at the pin jacks (plus or red jack is minus and vice versa). This can only be checked with

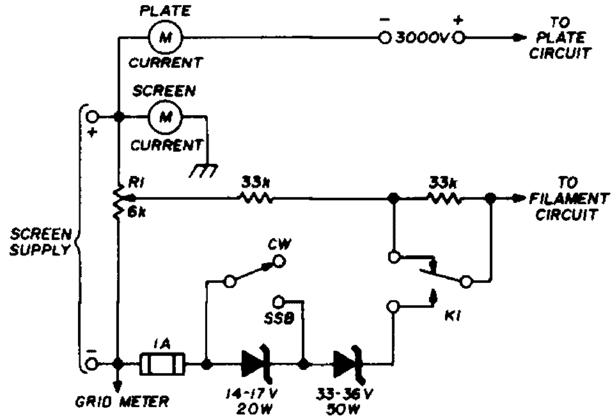


fig. 2. Improved metering and bias circuit for the five-band kilowatt linear. Zener diodes should be 50-volt units (50 watts) for CW and ssb operation.

another voltmeter or by checking a good diode that you know is properly marked. It's important to check zener diodes before installation because some manufacturers mark their product differently from others, and some diodes which appear on the surplus market are there because they were improperly marked in the first place.

John True, W4OQ
Great Falls, Virginia

AFSK generator

Dear HR:

Having built and tested the AFSK generator described in the December, 1973, issue, I found that the oscillator was very sluggish in starting (delays up to two minutes). Once started however, the unit functions beautifully.

The problem appears to be in the feedback loop of the oscillator, namely the crystal and capacitor C1. The feedback gain (β) is dependent upon the ratio of Cx/C1 where Cx is the capacitance of the crystal. This circuit seemed to need a little more feedback gain. Changing C1 to 510 pF cured the problem.

If the crystal is a type cut for parallel-resonant circuits (as are most of the military surplus crystals) the capacitance is different than that of a crystal cut for a series-resonant circuit, such as the circuit used in the AFSK generator.

David L. Chute, WA1NYL
Groton, Connecticut

the ham notebook

Heath HW-7 modifications

The Heath HW-7 QRP transceiver is a fine example of a compact rig for portable or home use. However, there are several minor modifications which will add to the ease of operations. These modifications are: an improved receiver blanking system, an adjustable sidetone volume level and the addition of a keyer.

The first modification is an improved circuit for cutting off the receiver during the keydown time. As is, a diode, CR2, is used to ground the high-gain audio amplifier's input during the key-down mode (see fig. 1). However, the amplifier's input is grounded during the time of key-down only, and not during key-up. Thus, an annoying clicking sound is heard as code is sent.

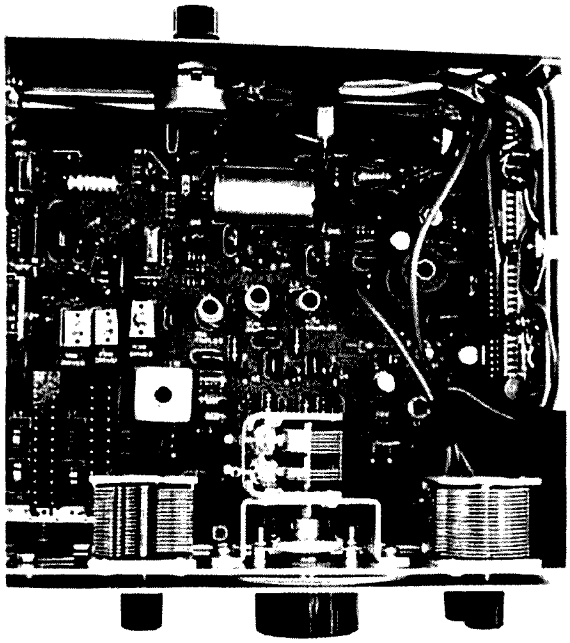


fig. 2. Location of the keyer board. Keyer controls are mounted on the rear panel. Note, at center, diode CR2 is removed from board. (Photo by WA5KVB)

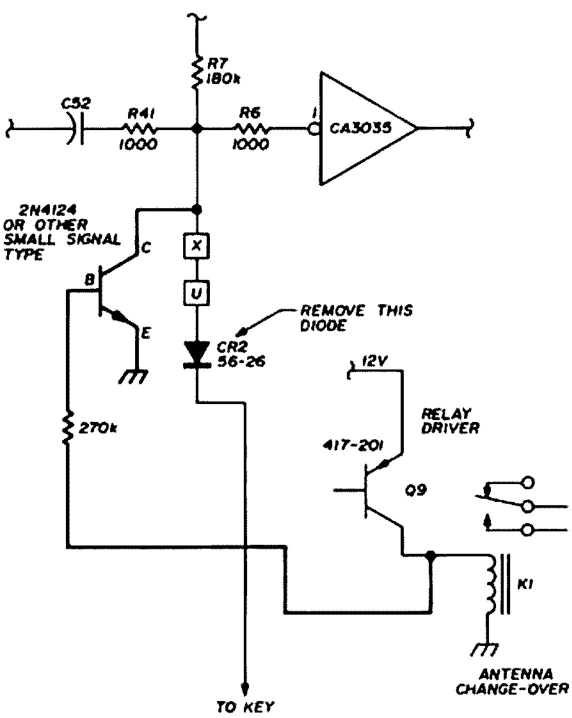


fig. 1. Schematic diagram of the HW-7 receiver blanking modification. The only two components needed, a small-signal transistor and a 270k resistor, are shown in bold.

An improved method of blanking the receiver during the transmit mode is to actually short the audio amplifier's input to ground during key-down *and* as long as the transceiver is in the transmit mode. The HW-7 has a time-delayed antenna transfer relay to switch between transmit and receive. By using the relay's coil

signal to switch a transistor gate, the receiver's audio amplifier's audio input will be grounded during the *total* transmit time.

The addition of a small signal transistor and a resistor is the only modification needed. **Fig. 1** shows the placement of the transistor switch in the circuit diagram. During the transmit mode, relay coil K1 is energized by 12-volts. This same voltage is used to turn on the added transistor, thus shorting the receiver's input. Diode CR2 should be removed as it is not needed.

sidetone volume

The sidetone amplitude is fixed in the HW-7. The addition of a 500k variable resistor in *series* with C45 (0.05 μF) will allow the adjustment of the sidetone's volume to a comfortable listening level. I used a miniature *Helipot* potentiometer, but the type is not critical. One lead of C45 is lifted from the foil; the potentiometer is connected in series with the capacitor. The sidetone volume is then easily adjusted by varying the added series resistor.

The addition of a keyer will be a welcome modification for the CW man. Rather than give a full description of the keyer design, the placement of the keyer in the HW-7 will be described (the keyer I used is a cmos version of the Accu-Keyer).[1] There is plenty of room inside the rig to allow for a great deal of flexibility. The keyer was placed in the rear, right-hand corner as shown in **fig. 2**. All of the keyer's controls were brought out to the rear panel since adjustments are seldom needed. A three-conductor phone jack replaces the phone jack originally used in the HW-7. This allows the use of a paddle-type key to be used externally. A tune pushbutton should also be mounted on the rear panel to facilitate tuning the rig. I recommend that a toggle switch not be used for the tuning control as the switch might be left on accidentally.

F.E. Hinkle, WA5KPG

1. James M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.

sensitive rf probe

To detect the very low-level rf signals you may find in receivers or low-level transmitter stages, it's necessary to use a rectifier that will respond to small signals and yet give a respectable output. There are two tricks that will help get this done.

One is to hand pick the signal diodes. Most amateurs have the needed test equipment on hand. All it takes is a source of dc readily varied from zero to 0.7 volts, a voltmeter you can read accurately in that range, a current-limiting resistor and a microammeter. Hook them up as shown in **fig. 3**. Take a handful of diodes such as you get for just about zero cost on surplus circuit boards and test each for its forward-conduction voltage. Select the germanium diodes with the lowest forward-conduction voltage.

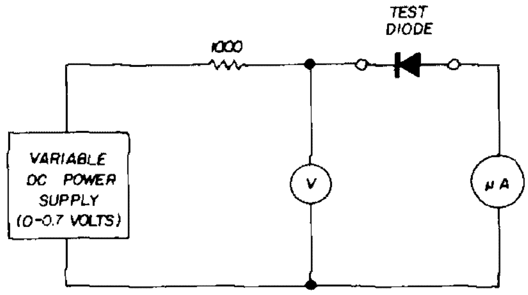


fig. 3. Circuit for testing diodes for use in the rf probe shown in fig. 4.

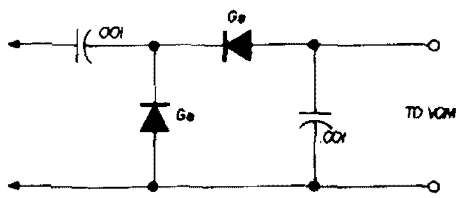


fig. 4. Sensitive rf probe uses selected germanium diodes in voltage-doubling circuit.

Then go back and rescreen that group for the ones with the least reverse-conduction current. Select two for use in the voltage-doubler rectifier shown in **fig. 4**.

This circuit, with selected germanium diodes, resulted in a change from zero to a full volt deflection on my fet-vom when used to pick up the output of a feeble rf signal generator.

Carl Drumeller, W5JJ

short circuits

L-network design

The radical sign in **equation 7** on page 27 of the February, 1974, issue should extend over the expression (X^2 + R^2) at the end of the line. This also applies to the formula for the constant k in the practical example on page 28.

ssb transceiver

In the article on the 40-meter ssb transmitter and receiver in the March, 1974, issue of *ham radio* the author used the wrong nomenclature for the Collins mechanical filter by calling it an FA21-7102. The correct nomenclature for this filter is F455FA21 (where F indicates a mechanical filter, 455 indicates a 455-kHz center frequency, FA is the case style [FA is used in the S-line] and 21 indicates a nominal 6-dB bandwidth of 2.1 kHz). The "7102" on the author's filter is simply a date code used by the manufacturer.

lowpass filters

The construction data for the 40-meter lowpass filter described in **fig.** 3 on page 39 of the March, 1974, issue is in error. L1 should be 15 turns number-16 on an Amidon T80-6 toroid (1.08 μH). L2 is 13 turns number-16 on an Amidon T80-6 core (0.76 μH). Insertion loss is approximately 0.14 dB.

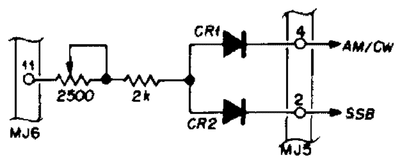
two-meter transverter

In the final amplifier schematic (**fig.** 5) for the two-meter transverter on page 12 of the February, 1974, issue of *ham radio* the symbol for the CTC byistor, BY1, is incorrect — the arrowhead should be at terminal S, the tail of the arrow at terminal I. Also, there should be a 68-ohm, 12-watt resistor in series with the line from terminal I of BY1 to the junction of the 0.1-μF capacitor and L104.

Yaesu sideband switching

A serious flaw appeared in the article on Yaesu sideband switching which appeared in the *ham notebook* column on page 57 of the December, 1973, issue. When the conversion is made as described, the USB and LSB are both on the same frequency on transmit and receive as claimed, but the tune and CW modes are shifted 3 kHz on receive but not changed on transmit, and the a-m mode is not changed on either transmit or receive! The problem is that pin 2 of MJ5 must be at ground potential to obtain the desired frequency shift — it is not in all cases with W2MUU's modification.

Fortunately, the solution is quite simple: simply add two silicon diodes as indicated in the schematic below.



The two diodes in this circuit operate as an OR gate so that whenever the emitter of Q6 (a-m/CW oscillator) or Q3 (USB oscillator) is grounded for operation, it also grounds the frequency-shift circuit.

cosmos electronic keyer

In the cosmos IC electronic keyer which was featured on page 6 of the June, 1974, issue, the circuit occasionally hangs up when first switched on. This can be corrected by removing pin 4 of U6A from ground and connecting it to pin 10 of U4B. Thus, when power is turned on, if the dot and dash generators both come on in the *on* state, the dash generator can now directly reset the dot generator, resulting in the emission of a single dash. After that the spurious state is permanently suppressed.

A second problem, where hang-up is induced by rapid deflection of the keyer paddle, is also eliminated by this modification. This problem is caused by the difference in propagation delays in the circuitry of the dot and dash generators (the delays in the dash generator are greater because the signal must travel through more gates). This problem is especially evident when going rapidly from dash to dot.

new products

portable general-coverage receiver

Barlow Wadley's new XCR-30 receiver is the first moderately-priced (around $260.00) all solid-state portable to feature direct frequency readout. Using a multiple heterodyne circuit (interpolation and crystal oscillators), the XCR-30 is a high sensitivity receiver designed to provide precision no-gap tuning from 500 kHz to 30 MHz. A 1-MHz crystal – in conjunction with the famous Wadley Loop circuit found in expensive ($2500 plus) Racal receivers – stabilizes the received frequency and eliminates drift. The tuned frequency is displayed mechanically as a composite function of two dials; the whole number (in MHz) is shown on one dial drum, and the decimal portion (in kHz) is shown on the second. Reception modes include a-m, CW and switchable single sideband.

The XCR-30 is metal-cased with external padding, not the usual plastic, and measures less than 300x200x100mm. Access to internal parts is through removal of the front or rear panel. The receiver has a built-in loudspeaker, but has facilities for headphones, external speaker, and 9-12 volt dc power source. In addition to the built-in, collapsible whip, an external antenna can be attached.

For more information on this exciting new receiver from South Africa, write to the American distributor, Gilfer Associates, Inc., Post Office Box 239, Park Ridge, New Jersey 07656, or use *check-off* on page 136.

hand-held two-meter transceiver

A new portable solid-state two-meter fm transceiver, designed to provide radio amateurs with reliable commercial quality performance at low cost, is now available from the Clegg Division of International Signal and Control Corporation. The two-watt, 5-channel unit features a unique battery saver design that results in less than 5-mA standby current drain while the high reliability battery offers up to 4 or 5 years of life under normal use.

The new HT-146 also features a single-conversion receiver, a monolithic crystal filter, and solid state T/R switching. Plug-in crystals make channel change fast and easy. Jacks for external microphone, speaker, and earphone are included along with BNC antenna connector and heliflex antenna. Accessories available include a tone encoder/decoder, microphone, leather case, earphone and an automatic battery charger.

The HT-146 hand-held transceiver is priced at $289.00. For additional information, write to Technical Literature Department, Clegg Division, International Signal and Control Corporation, 3050 Hempland Road, Lancaster, Pennsylvania 17601, or use *check-off* on page 136.

More Details? CHECK–OFF Page 136

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